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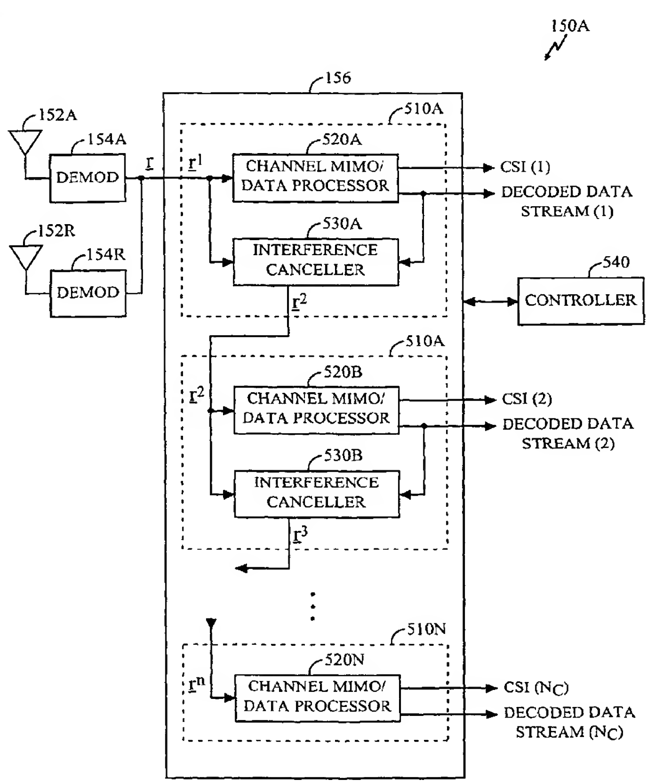
(54) Title: METHOD AND APPARATUS FOR PROCESSING DATA IN A MULTIPLE-INPUT MULTIPLE-OUTPUT (MIMO) COMMUNICATION SYSTEM UTILIZING CHANNEL STATE INFORMATION



(57) Abstract: Techniques to "successively" process received signals at a receiver unit in a MIMO system to recover transmitted data, and to "adaptively" process data at a transmitter unit based on channel state information available for the MIMO channel. A successive cancellation receiver processing technique is used to process the received signals and performs a number of iterations to provide decoded data streams. For each iteration, input (e.g., received) signals for the iteration are processed to provide one or more symbol streams. One of the symbol streams is selected and processed to provide a decoded data stream. The interference due to the decoded data stream is approximately removed (i.e., canceled) from the input signals provided to the next iteration. The channel characteristics are estimated and reported back to the transmitter system and used to adjust (i.e., adapt) the processing (e.g., coding, modulation, and so on) of data prior to transmission.

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METHOD AND APPARATUS FOR PROCESSING DATA IN A MULTIPLE-INPUT MULTIPLE-OUTPUT (MIMO) COMMUNICATION SYSTEM UTILIZING CHANNEL STATE INFORMATION

BACKGROUND

Field

[1001] The present invention relates generally to data communication, and more specifically to a novel and improved method and apparatus for processing data in a multiple-input multiple-output (MIMO) communication system utilizing channel state information to provide improved system performance.

Background

[1002] Wireless communication systems are widely deployed to provide various types of communication such as voice, data, and so on. These systems may be based on code division multiple access (CDMA), time division multiple access (TDMA), orthogonal frequency division multiplex (OFDM), or some other multiplexing techniques. OFDM systems may provide high performance for some channel environments.

[1003] In a terrestrial communication system (e.g., a cellular system, a broadcast system, a multi-channel multi-point distribution system (MMDS), and others), an RF modulated signal from a transmitter unit may reach a receiver unit via a number of transmission paths. The characteristics of the transmission paths typically vary over time due to a number of factors such as fading and multipath.

[1004] To provide diversity against deleterious path effects and improve performance, multiple transmit and receive antennas may be used for data transmission. If the transmission paths between the transmit and receive antennas are linearly independent (i.e., a transmission on one path is not formed as a linear combination of the transmissions on other paths), which is generally true to at least an extent, then the likelihood of correctly receiving a data transmission increases as the number of antennas increases. Generally, diversity increases and performance improves as the number of transmit and receive antennas increases.

[1005] A multiple-input multiple-output (MIMO) communication system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_C independent channels, with $N_C \leq \min \{N_T, N_R\}$. Each of the N_C independent channels is also referred to as a spatial subchannel of the MIMO channel and corresponds to a dimension. The MIMO system can provide improved performance (e.g., increased transmission capacity) if the additional dimensionalities created by the multiple transmit and receive antennas are utilized.

[1006] There is therefore a need in the art for techniques to process a data transmission at both the transmitter and receiver units to take advantage of the additional dimensionalities created by a MIMO system to provide improved system performance.

SUMMARY

[1007] Aspects of the invention provide techniques to process the received signals at a receiver unit in a multiple-input multiple-output (MIMO) system to recover the transmitted data, and to adjust the data processing at a transmitter unit based on estimated characteristics of a MIMO channel used for data transmission. In an aspect, a "successive cancellation" receiver processing technique (described below) is used to process the received signals. In another aspect, the channel characteristics are estimated and reported back to the transmitter system and used to adjust (i.e., adapt) the processing (e.g., coding, modulation, and so on) of data prior to transmission. Using a combination of the successive cancellation receiver processing technique and adaptive transmitter processing technique, high performance may be achieved for the MIMO system.

[1008] A specific embodiment of the invention provides a method for sending data from a transmitter unit to a receiver unit in a MIMO communication system. In accordance with the method, at the receiver unit, a number of signals are initially received via a number of receive antennas, with each received signal comprising a combination of one or more signals transmitted from the transmitter unit. The received signals are processed in accordance with a successive cancellation receiver processing technique to provide a number of decoded data streams, which are estimates of the data streams transmitted from the transmitter unit. Channel state information (CSI)

indicative of characteristics of a MIMO channel used to transmit the data steams are also determined and transmitted back to the transmitter unit. At the transmitter unit, each data stream is adaptively processed prior to transmission over the MIMO channel in accordance with the received CSI.

[1009] The successive cancellation receiver processing scheme typically performs a number of iterations to provide the decoded data streams, one iteration for each decoded data stream. For each iteration, a number of input signals for the iteration are processed in accordance with a particular linear or non-linear processing scheme to provide one or more symbol streams. One of the symbol streams is then selected and processed to provide a decoded data stream. A number of modified signals are also derived based on the input signals, with the modified signals having components due to the decoded data stream approximately removed (i.e., canceled). The input signals for a first iteration are the received signals and the input signals for each subsequent iteration are the modified signals from a preceding iteration.

[1010] Various linear and non-linear processing schemes may be used to process the input signals. For a non-dispersive channel (i.e., with flat fading), a channel correlation matrix inversion (CCMI) technique, a minimum mean square error (MMSE) technique, or some other techniques may be used. And for a time-dispersive channel (i.e., with frequency selective fading), an MMSE linear equalizer (MMSE-LE), a decision feedback equalizer (DFE), a maximum-likelihood sequence estimator (MLSE), or some other techniques may be used.

[1011] The available CSI may include, for example, the signal-to-noise-plus-interference (SNR) of each transmission channel to be used for data transmission. At the transmitter unit, the data for each transmission channel may be coded based on the CSI associated with that channel, and the coded data for each transmission channel may further be modulated in accordance with a modulation scheme selected based on the CSI.

[1012] The invention further provides methods, systems, and apparatus that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[1013] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[1014] FIG. 1 is a diagram of a multiple-input multiple-output (MIMO) communication system capable of implementing various aspects and embodiments of the invention;

[1015] FIG. 2 is a block diagram of an embodiment of a MIMO transmitter system capable of processing data for transmission based on the available CSI;

[1016] FIG. 3 is a block diagram of an embodiment of a MIMO transmitter system which utilizes orthogonal frequency division modulation (OFDM);

[1017] FIG. 4 is a flow diagram illustrating a successive cancellation receiver processing technique to process N_R received signals to recover N_T transmitted signals;

[1018] FIG. 5 is a block diagram of a receiver system capable of implementing various aspects and embodiments of the invention;

[1019] FIGS. 6A, 6B, and 6C are block diagrams of three channel MIMO/data processors, which are capable of implementing a CCMI technique, a MMSE technique, and a DFE technique, respectively;

[1020] FIG. 7 is a block diagram of an embodiment of a receive (RX) data processor;

[1021] FIG. 8 is a block diagram of an interference canceller; and

[1022] FIGS. 9A, 9B, and 9C are plots that illustrate the performance for various receiver and transmitter processing schemes.

DETAILED DESCRIPTION

[1023] FIG. 1 is a diagram of a multiple-input multiple-output (MIMO) communication system 100 capable of implementing various aspects and embodiments of the invention. System 100 includes a first system 110 in communication with a second system 150. System 100 can be operated to employ a combination of antenna, frequency, and temporal diversity (described below) to increase spectral efficiency, improve performance, and enhance flexibility. In an aspect, system 150 can be operated to determine the characteristics of a MIMO channel and to report channel state

information (CSI) indicative of the channel characteristics that have been determined in this way back to system 110, and system 110 can be operated to adjust the processing (e.g., encoding and modulation) of data prior to transmission based on the available CSI. In another aspect, system 150 can be operated to process the data transmission from system 110 in a manner to provide high performance, as described in further detail below.

[1024] At system 110, a data source 112 provides data (i.e., information bits) to a transmit (TX) data processor 114, which encodes the data in accordance with a particular encoding scheme, interleaves (i.e., reorders) the encoded data based on a particular interleaving scheme, and maps the interleaved bits into modulation symbols for one or more transmission channels used for transmitting the data. The encoding increases the reliability of the data transmission. The interleaving provides time diversity for the coded bits, permits the data to be transmitted based on an average signal-to-noise-plus-interference-ratio (SNR) for the transmission channels used for the data transmission, combats fading, and further removes correlation between coded bits used to form each modulation symbol. The interleaving may further provide frequency diversity if the coded bits are transmitted over multiple frequency subchannels. In an aspect, the encoding, interleaving, and symbol mapping (or a combination thereof) are performed based on the CSI available to system 110, as indicated in FIG. 1.

[1025] The encoding, interleaving, and symbol mapping at transmitter system 110 can be performed based on numerous schemes. One specific scheme is described in U.S Patent Application Serial No. 09/776,075, entitled "CODING SCHEME FOR A WIRELESS COMMUNICATION SYSTEM," filed February 1, 2001, assigned to the assignee of the present application and incorporated herein by reference. Another scheme is described in further detail below.

[1026] MIMO system 100 employs multiple antennas at both the transmit and receive ends of the communication link. These transmit and receive antennas may be used to provide various forms of spatial diversity (i.e., antenna diversity), including transmit diversity and receive diversity. Spatial diversity is characterized by the use of multiple transmit antennas and one or more receive antennas. Transmit diversity is characterized by the transmission of data over multiple transmit antennas. Typically, additional processing is performed on the data transmitted from the transmit antennas to achieved the desired diversity. For example, the data transmitted from different

transmit antennas may be delayed or reordered in time, coded and interleaved across the available transmit antennas, and so on. Receive diversity is characterized by the reception of the transmitted signals on multiple receive antennas, and diversity is achieved by simply receiving the signals via different signal paths.

[1027] System 100 may be operated in a number of different communication modes, with each communication mode employing antenna, frequency, or temporal diversity, or a combination thereof. The communication modes may include, for example, a "diversity" communication mode and a "MIMO" communication mode. The diversity communication mode employs diversity to improve the reliability of the communication link. In a common application of the diversity communication mode, which is also referred to as a "pure" diversity communication mode, data is transmitted from all available transmit antennas to a recipient receiver system. The pure diversity communications mode may be used in instances where the data rate requirements are low or when the SNR is low, or when both are true. The MIMO communication mode employs antenna diversity at both ends of the communication link (i.e., multiple transmit antennas and multiple receive antennas) and is generally used to both improve the reliability and increase the capacity of the communication link. The MIMO communication mode may further employ frequency and/or temporal diversity in combination with the antenna diversity.

[1028] System 100 may utilize orthogonal frequency division modulation (OFDM), which effectively partitions the operating frequency band into a number of (N_L) frequency subchannels (i.e., frequency bins). At each time slot (i.e., a particular time interval that may be dependent on the bandwidth of the frequency subchannel), a modulation symbol may be transmitted on each of the N_L frequency subchannels.

[1029] System 100 may be operated to transmit data via a number of transmission channels. As noted above, a MIMO channel may be decomposed into N_C independent channels, with $N_C \leq \min \{N_T, N_R\}$. Each of the N_C independent channels is also referred to as a spatial subchannel of the MIMO channel. For a MIMO system not utilizing OFDM, there is typically only one frequency subchannel and each spatial subchannel may be referred to as a "transmission channel". For a MIMO system utilizing OFDM, each spatial subchannel of each frequency subchannel may be referred to as a transmission channel.

[1030] A MIMO system can provide improved performance if the additional dimensionalities created by the multiple transmit and receive antennas are utilized. While this does not necessarily require knowledge of CSI at the transmitter, increased system efficiency and performance are possible when the transmitter is equipped with CSI, which is descriptive of the transmission characteristics from the transmit antennas to the receive antennas. The processing of data at the transmitter prior to transmission is dependent on whether or not CSI is available.

[1031] The available CSI may comprise, for example, the signal-to-noise-plus-interference-ratio (SNR) of each transmission channel (i.e., the SNR for each spatial subchannel for a MIMO system without OFDM, or the SNR for each spatial subchannel of each frequency subchannel for a MIMO system with OFDM). In this case, data may be adaptively processed at the transmitter (e.g., by selecting the proper coding and modulation scheme) for each transmission channel based on the channel's SNR.

[1032] For a MIMO system not employing OFDM, TX MIMO processor 120 receives and demultiplexes the modulation symbols from TX data processor 114 and provides a stream of modulation symbols for each transmit antenna, one modulation symbol per time slot. And for a MIMO system employing OFDM, TX MIMO processor 120 provides a stream of modulation symbol vectors for each transmit antenna, with each vector including N_L modulation symbols for the N_L frequency subchannels for a given time slot. Each stream of modulation symbols or modulation symbol vectors is received and modulated by a respective modulator (MOD) 122, and transmitted via an associated antenna 124.

[1033] At receiver system 150, a number of receive antennas 152 receive the transmitted signals and provide the received signals to respective demodulators (DEMOD) 154. Each demodulator 154 performs processing complementary to that performed at modulator 122. The modulation symbols from all demodulators 154 are provided to a receive (RX) MIMO/data processor 156 and processed to recover the transmitted data streams. RX MIMO/data processor 156 performs processing complementary to that performed by TX data processor 114 and TX MIMO processor 120 and provides decoded data to a data sink 160. The processing by receiver system 150 is described in further detail below.

[1034] The spatial subchannels of a MIMO system (or more generally, the transmission channels in a MIMO system with or without OFDM) typically experience

different link conditions (e.g., different fading and multipath effects) and may achieve different SNR. Consequently, the capacity of the transmission channels may be different from channel to channel. This capacity may be quantified by the information bit rate (i.e., the number of information bits per modulation symbol) that may be transmitted on each transmission channel for a particular level of performance (e.g., a particular bit error rate (BER) or packet error rate (PER)). Moreover, the link conditions typically vary with time. As a result, the supported information bit rates for the transmission channels also vary with time. To more fully utilize the capacity of the transmission channels, CSI descriptive of the link conditions may be determined (typically at the receiver unit) and provided to the transmitter unit so that the processing can be adjusted (or adapted) accordingly. The CSI may comprise any type of information that is indicative of the characteristics of the communication link and may be reported via various mechanisms, as described in further detail below. For simplicity, various aspects and embodiments of the invention are described below wherein the CSI comprises SNR. Techniques to determine and utilize CSI to provide improved system performance are described below.

MIMO Transmitter System with CSI Processing

[1035] FIG. 2 is a block diagram of an embodiment of a MIMO transmitter system 110a, which does not utilize OFDM but is capable of adjusting its processing based on CSI available to the transmitter system (e.g., as reported by receiver system 150). Transmitter system 110a is one embodiment of the transmitter portion of system 110 in FIG. 1. System 110a includes (1) a TX data processor 114a that receives and processes information bits to provide modulation symbols and (2) a TX MIMO processor 120a that demultiplexes the modulation symbols for the N_T transmit antennas.

[1036] In the specific embodiment shown in FIG. 2, TX data processor 114a includes a demultiplexer 208 coupled to a number of channel data processors 210, one processor for each of the N_C transmission channels. Demultiplexer 208 receives and demultiplexes the aggregate information bits into a number of (up to N_C) data streams, one data stream for each of the transmission channels to be used for data transmission. Each data stream is provided to a respective channel data processor 210.

[1037] In the embodiment shown in FIG. 2, each channel data processor 210 includes an encoder 212, a channel interleaver 214, and a symbol mapping element 216.

Encoder 212 receives and encodes the information bits in the received data stream in accordance with a particular encoding scheme to provide coded bits. Channel interleaver 214 interleaves the coded bits based on a particular interleaving scheme to provide diversity. And symbol mapping element 216 maps the interleaved bits into modulation symbols for the transmission channel used for transmitting the data stream.

[1038] Pilot data (e.g., data of known pattern) may also be encoded and multiplexed with the processed information bits. The processed pilot data may be transmitted (e.g., in a time division multiplexed (TDM) manner) in all or a subset of the transmission channels used to transmit the information bits. The pilot data may be used at the receiver to perform channel estimation, as described below.

[1039] As shown in FIG. 2, the data encoding, interleaving, and modulation (or a combination thereof) may be adjusted based on the available CSI (e.g., as reported by receiver system 150). In one coding and modulation scheme, adaptive encoding is achieved by using a fixed base code (e.g., a rate 1/3 Turbo code) and adjusting the puncturing to achieve the desired code rate, as supported by the SNR of the transmission channel used to transmit the data. For this scheme, the puncturing may be performed after the channel interleaving. In another coding and modulation scheme, different coding schemes may be used based on the reported CSI. For example, each of the data streams may be coded with an independent code. With this scheme, a "successive cancellation" receiver processing scheme may be used to detect and decode the data streams to derive a more reliable estimate of the transmitted data streams, as described in further detail below.

[1040] Symbol mapping element 216 can be designed to group sets of interleaved bits to form non-binary symbols, and to map each non-binary symbol into a point in a signal constellation corresponding to a particular modulation scheme (e.g., QPSK, M-PSK, M-QAM, or some other scheme) selected for the transmission channel. Each mapped signal point corresponds to a modulation symbol.

[1041] The number of information bits that may be transmitted for each modulation symbol for a particular level of performance (e.g., one percent PER) is dependent on the SNR of the transmission channel. Thus, the coding and modulation scheme for each transmission channel may be selected based on the available CSI. The channel interleaving may also be adjusted based on the available CSI.

[1042] Table 1 lists various combinations of coding rate and modulation scheme that may be used for a number of SNR ranges. The supported bit rate for each transmission channel may be achieved using any one of a number of possible combinations of coding rate and modulation scheme. For example, one information bit per modulation symbol may be achieved using (1) a coding rate of 1/2 and QPSK modulation, (2) a coding rate of 1/3 and 8-PSK modulation, (3) a coding rate of 1/4 and 16-QAM, or some other combination of coding rate and modulation scheme. In Table 1, QPSK, 16-QAM, and 64-QAM are used for the listed SNR ranges. Other modulation schemes such as 8-PSK, 32-QAM, 128-QAM, and so on, may also be used and are within the scope of the invention.

Table 1

SNR Range	# of Information Bits/Symbol	Modulation Symbol	# of Coded Bits/Symbol	Coding Rate
1.5 – 4.4	1	QPSK	2	1/2
4.4 – 6.4	1.5	QPSK	2	3/4
6.4 – 8.35	2	16-QAM	4	1/2
8.35 – 10.4	2.5	16-QAM	4	5/8
10.4 – 12.3	3	16-QAM	4	3/4
12.3 – 14.15	3.5	64-QAM	6	7/12
14.15 – 15.55	4	64-QAM	6	2/3
15.55 – 17.35	4.5	64-QAM	6	3/4
> 17.35	5	64-QAM	6	5/6

[1043] The modulation symbols from TX data processor 114a are provided to a TX MIMO processor 120a, which is one embodiment of TX MIMO processor 120 in FIG. 1. Within TX MIMO processor 120a, a demultiplexer 222 receives (up to) N_C modulation symbol streams from N_C channel data processors 210 and demultiplexes the received modulation symbols into a number of (N_T) modulation symbol streams, one stream for each antenna used to transmit the modulation symbols. Each modulation symbol stream is provided to a respective modulator 122. Each modulator 122 converts the modulation symbols into an analog signal, and further amplifies, filters, quadrature modulates, and upconverts the signal to generate a modulated signal suitable for transmission over the wireless link.

MIMO Transmitter System with OFDM

[1044] FIG. 3 is a block diagram of an embodiment of a MIMO transmitter system 110c, which utilizes OFDM and is capable of adjusting its processing based on the available CSI. Within a TX data processor 114c, the information bits to be transmitted are demultiplexed into a number of (up to N_L) frequency subchannel data streams, one stream for each of the frequency subchannels to be used for the data transmission. Each frequency subchannel data stream is provided to a respective frequency subchannel data processor 310.

[1045] Each data processor 310 processes data for a respective frequency subchannel of the OFDM system. Each data processor 310 may be implemented similar to TX data processor 114a shown in FIG. 2. For this design, data processor 310 includes a demultiplexer that demultiplexes the frequency subchannel data stream into a number of data substreams, one substream for each spatial subchannel used for the frequency subchannel. Each data substream is then encoded, interleaved, and symbol mapped by a respective channel data processor to generate modulation symbols for that particular transmission channel (i.e., that spatial subchannel of that frequency subchannel). The coding and modulation for each transmission channel may be adjusted based on the available CSI (e.g., reported by the receiver system). Each frequency subchannel data processor 310 thus provides (up to) N_C modulation symbol streams for (up to) N_C spatial subchannels.

[1046] For a MIMO system utilizing OFDM, the modulation symbols may be transmitted on multiple frequency subchannels and from multiple transmit antennas. Within a MIMO processor 120c, the N_C modulation symbol streams from each data processor 310 are provided to a respective channel MIMO processor 322, which processes the received modulation symbols based on the available CSI.

[1047] Each channel MIMO processor 322 demultiplexes the N_C modulation symbols for each time slot into N_T modulation symbols for the N_T transmit antennas. Each combiner 324 receives the modulation symbols for up to N_L frequency subchannels, combines the symbols for each time slot into a modulation symbol vector V, and provides the modulation symbol vector to the next processing stage (i.e., a respective modulator 122).

[1048] MIMO processor 120c thus receives and processes the modulation symbols to provide N_T modulation symbol vectors, V_1 through V_{Nt}, one modulation symbol vector for each transmit antenna. Each modulation symbol vector V covers a single time slot, and each element of the modulation symbol vector V is associated with a specific frequency subchannel having a unique subcarrier on which the modulation symbol is conveyed.

[1049] FIG. 3 also shows an embodiment of modulator 122 for OFDM. The modulation symbol vectors V_1 through V_{Nt} from MIMO processor 120c are provided to modulators 122a through 122t, respectively. In the embodiment shown in FIG. 3, each modulator 122 includes an inverse Fast Fourier Transform (IFFT) 320, cycle prefix generator 322, and an upconverter 324.

[1050] IFFT 320 converts each received modulation symbol vector into its time-domain representation (which is referred to as an OFDM symbol) using IFFT. IFFT 320 can be designed to perform the IFFT on any number of frequency subchannels (e.g., 8, 16, 32, and so on). In an embodiment, for each modulation symbol vector converted to an OFDM symbol, cycle prefix generator 322 repeats a portion of the time-domain representation of the OFDM symbol to form a "transmission symbol" for a specific transmit antenna. The cyclic prefix insures that the transmission symbol retains its orthogonal properties in the presence of multipath delay spread, thereby improving performance against deleterious multipath effects. The implementation of IFFT 320 and cycle prefix generator 322 is known in the art and not described in detail herein.

[1051] The time-domain representations from each cyclic prefix generator 322 (i.e., the transmission symbols for each antenna) are then processed (e.g., converted into an analog signal, modulated, amplified, and filtered) by upconverter 324 to generate a modulated signal, which is then transmitted from the respective antenna 124.

[1052] OFDM modulation is described in further detail in a paper entitled "Multicarrier Modulation for Data Transmission : An Idea Whose Time Has Come," by John A.C. Bingham, IEEE Communications Magazine, May 1990, which is incorporated herein by reference.

[1053] FIGS. 2 and 3 show two designs of a MIMO transmitter capable of implementing various aspects of the invention. Other transmitter designs may also be implemented and are within the scope of the invention. Some of these transmitter designs are described in further detail in U.S Patent Application Serial No. 09/532,492,

entitled "HIGH EFFICIENCY, HIGH PERFORMANCE COMMUNICATIONS SYSTEM EMPLOYING MULTI-CARRIER MODULATION," filed March 22, 2000, the aforementioned U.S Patent Application Serial No. 09/776,075, and U.S Patent Application Serial No. 09/826,481 "METHOD AND APPARATUS FOR UTILIZING CHANNEL STATE INFORMATION IN A WIRELESS COMMUNICATION SYSTEM," filed March 23, 2001, all assigned to the assignee of the present application and incorporated herein by reference. These patent applications describe MIMO processing and CSI processing in further detail.

[1054] In general, transmitter system 110 codes and modulates data for each transmission channel based on information descriptive of that channel's transmission capability. This information is typically in the form of CSI. The CSI for the transmission channels used for data transmission is typically determined at the receiver system and reported back to the transmitter system, which then uses the information to adjust the coding and modulation accordingly. The techniques described herein are applicable for multiple parallel transmission channels supported by MIMO, OFDM, or any other communication scheme (e.g., a CDMA scheme) capable of supporting multiple parallel transmission channels.

MIMO Receiver System

[1055] Aspects of the invention provide techniques to (1) process the received signals at a receiver system in a MIMO system based on a successive cancellation receiver processing scheme to recover the transmitted data, and (2) adjust the data processing at a transmitter system based on estimated characteristics of the MIMO channel. In an aspect, the successive cancellation receiver processing technique (described below) is used to process the received signals. In another aspect, the channel characteristics are estimated at the receiver system and reported back to the transmitter system, which uses the information to adjust (i.e., adapt) the data processing (e.g., coding, modulation, and so on). Using a combination of the successive cancellation receiver processing technique and adaptive transmitter processing technique, high performance may be achieved for the MIMO system.

[1056] FIG. 4 is a flow diagram illustrating the successive cancellation receiver processing technique to process N_R received signals to recover N_T transmitted signals. For simplicity, the following description for FIG. 4 assumes that (1) the number of

transmission channels (i.e., spatial subchannels for a MIMO system not utilizing OFDM) is equal to the number of transmit antenna (i.e., $N_C = N_T$) and (2) one independent data stream is transmitted from each transmit antenna.

[1057] Initially, the receiver system performs linear and/or non-linear space-processing on the N_R received signals to attempt to separate the multiple transmitted signals included in the received signals, at step 412. Linear spatial processing may be performed on the received signals if the MIMO channel is "non-dispersive" (i.e., frequency non-selective or flat fading).It may also be necessary or desirable to perform additional linear or non-linear temporal processing (i.e., equalization) on the received signals if the MIMO channel is "time-dispersive" (i.e., frequency selective fading). The spatial processing may be based on a channel correlation matrix inversion (CCMI) technique, a minimum mean square error (MMSE) technique, or some other technique. The space-time processing may be based on an MMSE linear equalizer (MMSE-LE), a decision feedback equalizer (DFE), a maximum-likelihood sequence estimator (MLSE), or some other technique. Some of these spatial and space-time processing techniques are described in further detail below. The amount of achievable signal separation is dependent on the amount of correlation between the transmitted signals, and greater signal separation may be obtained if the transmitted signals are less correlated.

[1058] The initial spatial or space-time processing step provides N_T "post-processed" signals that are estimates of the N_T transmitted signals. The SNRs for the N_T post-processed signals are then determined, at step 414. The SNR may be estimated as described in further detail below. In one embodiment, the SNRs are ranked in order from highest to lowest SNR, and the post-processed signal having the highest SNR is selected and further processed (i.e., "detected") to obtain a decoded data stream, at step 416. The detection typically includes demodulating, deinterleaving, and decoding the selected post-processed signal. The decoded data stream is an estimate of the data stream transmitted on the transmitted signal being recovered in this iteration. The particular post-processed signal to be detected may also be selected based on some other scheme (e.g., the particular signal may be specifically identified by the transmitter system).

[1059] At step 418, a determination is made whether or not all transmitted signals have been recovered. If all transmitted signals have been recovered, then the receiver processing terminates. Otherwise, the interference due to the decoded data stream is

removed from the received signals to generate "modified" signals for the next iteration to recover the next transmitted signal.

[1060] At step 420, the decoded data stream is used to form an estimate of the interference presented by the transmitted signal corresponding to the decoded data stream on each of the received signals. The interference can be estimated by first re-encoding the decoded data stream, interleaving the re-encoded data, and symbol mapping the interleaved data (using the same coding, interleaving, and modulation schemes used at the transmitter for this data stream) to obtain a stream of "remodulated" symbols. The remodulated symbol stream is an estimate of the modulation symbol stream previously transmitted from one of the N_T transmit antennas and received by the N_R received antennas. Thus, the remodulated symbol stream is convolved by each of N_R elements in an estimated channel response vector $\underline{\mathbf{h}}_i$ to derive N_R interference signals due to the recovered transmitted signal. The vector $\underline{\mathbf{h}}_i$ is a particular column of a $(N_R \times N_T)$ channel coefficient matrix $\mathbf{H}$, which represents an estimate of the MIMO channel response for the N_T transmit antennas and N_R receive antennas at a specific time and which may be derived based on pilot signals transmitted along with the data. The N_R interference signals are then subtracted from the N_R corresponding received signals to derive N_R modified signals, at step 422. These modified signals represent the signals at the received antennas if the components due to the decoded data stream had not been transmitted (i.e., assuming that the interference cancellation was effectively performed).

[1061] The processing performed in steps 412 through 416 is then repeated on the N_R modified signals (instead of the N_R received signals) to recover another transmitted signal. Steps 412 through 416 are thus repeated for each transmitted signal to be recovered, and steps 420 and 422 are performed if there is another transmitted signal to be recovered.

[1062] The successive cancellation receiver processing technique thus performs a number of iterations, one iteration for each transmitted signal to be recovered. Each iteration (except for the last) performs a two-part processing to recover one of the transmitted signals and to generate the modified signals for the next iteration. In the first part, spatial processing or space-time processing is performed on the N_R received signals to provide N_R post-processed signals, and one of the post-processed signals is detected to recover the data stream corresponding to this transmitted signal. In the second part (which needs not be performed for the last iteration), interference due to the

decoded data stream is canceled from the received signals to derive modified signals having the recovered component removed.

Initially, the input signals for the first iteration are the received signals, which may be expressed as:

$$\underline{\mathbf{r}}^1 = \underline{\mathbf{r}} = \begin{bmatrix} r_1 \\ r_2 \\ \mathrm{M} \\ r_{N_R} \end{bmatrix} , \qquad \text{Eq (1)}$$

where $\underline{\mathbf{r}}$ is the vector of N_R received signals and $\underline{\mathbf{r}}^1$ is the vector of N_R input signals for the first iteration of the successive cancellation receiver processing scheme. These input signals are linearly or non-linearly processed to provide post-processed signals, which may be expressed as:

$$\underline{\mathbf{x}}^1 = \begin{bmatrix} x_1^1 \\ x_2^1 \\ \mathrm{M} \\ x_{N_T}^1 \end{bmatrix} \qquad \text{Eq (2)}$$

where $\underline{\mathbf{x}}^1$ is the vector of N_R post-processed signals from the first iteration. The SNR of the post-processed signals may be estimated, which may be expressed as:

$$\underline{\gamma}^1 = [\gamma_1^1,\ \gamma_2^1,\ \ldots,\ \gamma_{N_T}^1] \ . \qquad \text{Eq (3)}$$

[1063] One of the post-processed signals is selected for further processing (e.g., the post-processed signal with the highest SNR) to provide a decoded data stream. This decoded data stream is then used to estimate the interference $\underline{\mathbf{i}}^1$ generated by the recovered signal, which may be expressed as:

$$\hat{\underline{\mathbf{i}}}^1 = \begin{bmatrix} \hat{i}_1^1 \\ \hat{i}_2^1 \\ \mathrm{M} \\ \hat{i}_{N_R}^1 \end{bmatrix} . \qquad \text{Eq (4)}$$

The interference $\hat{\underline{\mathbf{i}}}^1$ is then subtracted from the input signal vector $\underline{\mathbf{r}}^1$ for this iteration to derive modified signals that comprise the input signal vector $\underline{\mathbf{r}}^2$ for the next iteration. The interference cancellation may be expressed as:

$$\underline{\mathbf{r}}^2 = \underline{\mathbf{r}}^1 - \hat{\underline{\mathbf{i}}}^1 = \begin{bmatrix} r_1^1 - \hat{i}_1^1 \\ r_2^1 - \hat{i}_2^1 \\ \vdots \\ r_{N_R}^1 - \hat{i}_{N_R}^1 \end{bmatrix} \qquad \text{Eq (5)}$$

[1064] The same process is then repeated for the next iteration, with the vector $\underline{\mathbf{r}}^2$ comprising the input signals for this iteration.

[1065] With the successive cancellation receiver processing scheme, one transmitted signal is recovered for each iteration, and the SNR for the i-th transmitted signal recovered in the k-th iteration, γ_i^k, may be provided as the CSI for the transmission channel used to transmit this recovered signal. As an example, if the first post-processed signal x_1^1 is recovered in the first iteration, the second post-processed signal x_2^2 is recovered in the second iteration, and so on, and the N_T-th post-processed signal $x_{N_T}^{N_T}$ is recovered in the last iteration, then the CSI that may be reported for these recovered signals may be expressed as: $\underline{\gamma} = [\gamma_1^1,\ \gamma_2^2,\ \ldots,\ \gamma_{N_T}^{N_T}]$.

[1066] Using the successive cancellation receiver processing technique, the original N_R received signals are thus successively processed to recover one transmitted signal at a time. Moreover, each recovered transmitted signal is removed (i.e., canceled) from the received signals prior to the processing to recover the next transmitted signal. If the transmitted data streams can be decoded without error (or with minimal errors) and if the channel response estimate is reasonably accurately, then the cancellation of interference due to previously recovered transmitted signals from the received signals is effective. The interference cancellation typically improves the SNR of each transmitted signal to be subsequently recovered. In this way, higher performance may be achieved for all transmitted signals (possibly except for the first transmitted signal to be recovered).

[1067] The possible improvement in SNR for the recovered transmitted signals using the successive cancellation receiver processing technique may be illustrated by an example. In this example, a pair of cross-polarized antennas is employed at both the

transmitter and receiver, the MIMO channel is line-of-sight, and four independent data streams are transmitted on the vertical and horizontal components of the pair of cross-polarized transmit antennas. For simplicity, it is assumed that the cross-polarization isolation is perfect so that the vertical and horizontal components do not interfere with one another at the receiver.

[1068] The receiver initially receives four signals on the vertical and horizontal components of the pair of cross-polarized received antennas and processes these four received signals. The received signals on the vertical elements of the cross-polarized antennas are highly correlated, and the received signals on the horizontal elements are similarly highly correlated.

[1069] When there is a strong linear dependence between two or more transmit-receive antenna pairs composing the MIMO channel, the ability to null interference is compromised. In this case, the linear spatial processing will be unsuccessful at separating the four independent data streams transmitted on the vertical and horizontal components of the pair of cross-polarized antennas. Specifically, the vertical component on each cross-polarized transmit antenna interferes with the vertical component on the other cross-polarized transmit antenna, and similar interference is experienced on the horizontal component. Thus, the resulting SNR for each of the four transmitted signals will be poor due to the correlated interference from the other antenna with the same polarization. As a result, the capacity of the transmitted signals based only on linear spatial processing will be severely constrained by the correlated interference signal.

[1070] When the eigenmodes for this example MIMO channel are examined, it can be seen that there are only two non-zero eigenmodes (i.e., the vertical and horizontal polarizations). A "full-CSI" processing scheme would then transmit only two independent data streams using these two eigenmodes. The capacity achieved in this case can be expressed as:

$$Capacity = 2 \cdot \log_2(1 + \lambda_i / \sigma^2) \ ,$$

where λ_i / σ^2 is the ratio of received signal power to thermal noise power for the i-th eigenmode. Thus, the capacity of the full-CSI processing scheme for this example MIMO channel is identical to the capacity of two parallel additive white Gaussian noise (AWGN) channels, each having an SNR given by λ_i / σ^2.

[1071] With the successive cancellation receiver processing technique, the linear spatial processing performed in step 412 initially results in the SNR for each of the four transmitted signals being 0 dB or less (due to the noise plus interference from the other transmitted signal on the same polarization). The overall capacity would be poor if no additional receiver processing is performed.

[1072] However, by applying successive spatial processing and interference cancellation, the SNR of subsequently recovered transmitted signals can be improved. For example, the first transmitted signal to be recovered may be the vertical polarization from the first cross-polarized transmit antenna. If it is assumed that the interference cancellation is effectively performed (i.e., zero or minimal decision errors and accurate channel estimates), then this signal no longer (or minimally) interferes with the remaining three (not yet recovered) transmitted signals. Removing this vertical polarization interference improves the SNR on the other not yet recovered signal transmitted on the vertical polarization. The cross-polarization isolation was assumed to be perfect for this simple example, and the two signals transmitted on the horizontal polarization do not interference with the signals transmitted on the vertical polarization. Thus, with effective interference cancellation, the signal transmitted on the vertical polarization of the second cross-polarized transmit antenna may be recovered at an SNR that is (theoretically) limited by thermal noise power.

[1073] In the above example, removing the interference from the vertical polarization does not impact the SNR of the two signals transmitted on the horizontal polarizations. Thus, the successive spatial processing and interference cancellation are similarly applied for the two signals transmitted on the horizontal polarization. This results in the first recovered signal on the horizontal polarization having a low SNR and the second recovered signal on the horizontal polarization having an SNR that is also (theoretically) limited by thermal noise.

[1074] As a result of performing successive spatial processing and interference cancellation, the two transmitted signals with low SNR contribute little to the total capacity, but the two transmitted signals with high SNR contribute in a significant manner to the total capacity.

Non-Dispersive and Dispersive Channels

[1075] Different receive and (possibly) transmit processing schemes may be used depending on the characteristics of the MIMO channel, which may be characterized as either non-dispersive or dispersive. A non-dispersive MIMO channel experiences flat fading (i.e., frequency non-selective fading), which may be more likely when the system bandwidth is narrow. A dispersive MIMO channel experiences frequency non-selective fading (e.g., different amount of attenuation across the system bandwidth), which may be more likely when the system bandwidth is wide and for certain operating conditions and environments. The successive cancellation receiver processing technique can be advantageously used for both non-dispersive and dispersive MIMO channels.

[1076] For a non-dispersive MIMO channel, linear spatial processing techniques such as CCMI and MMSE may be used to process the received signals prior to demodulation and decoding. These linear spatial processing techniques may be employed at the receiver to null out the undesired signals, or to maximize the received signal-to-interference-plus-noise ratio of each of the constituent signals in the presence of noise and interference from the other signals. The ability to effectively null undesired signals or optimize the signal-to-interference-plus-noise ratios depends upon the correlation in the channel coefficient matrix $\underline{\mathbf{H}}$ that describes the channel response between the transmit and receive antennas. The successive cancellation receiver processing technique (e.g., with CCMI or MMSE) can be advantageously used for non-dispersive MIMO channel.

[1077] For a dispersive MIMO channel, time dispersion in the channel introduces inter-symbol interference (ISI). To improve performance, a wideband receiver attempting to recover a particular transmitted data stream would need to ameliorate both "crosstalk" from the other transmitted signals as well as inter-symbol interference from all of the transmitted signals. The successive cancellation receiver processing technique can be extended to handle dispersive MIMO channel. To deal with crosstalk and inter-symbol interference, the spatial processing in a narrowband receiver (which handles crosstalk well but does not effectively deal with inter-symbol interference) may be replaced with space-time processing in the wideband receiver. In the wideband receiver, the successive cancellation receiver processing technique may be employed in similar manner as that described above for FIG. 4. However, the spatial processing performed in step 412 is replaced with space-time processing.

[1078] In one embodiment, a MMSE linear equalizer (MMSE-LE) may be used for the space-time processing in a wideband receiver. Using the MMSE-LE technique, the space-time processing assumes similar form as the spatial processing for the narrowband channel. However, each "filter tap" in the spatial processor includes more than one tap, as described in further detail below. The MMSE-LE technique is most effective for use in space-time processing when the channel estimates (i.e., the channel coefficient matrix $\underline{\mathbf{H}}$) are accurate.

[1079] In another embodiment, a decision feedback equalizer (DFE) may be used for the space-time processing at the wideband receiver. The DFE is a non-linear equalizer that is effective for channels with severe amplitude distortion and uses decision feedback to cancel interference from symbols that have already been detected. If the data stream can be decoded without errors (or with minimal errors), then the inter-symbol interference generated by the modulation symbols corresponding to the decoded data bits may be effectively canceled.

[1080] In yet another embodiment, a maximum-likelihood sequence estimator (MLSE) may be used for the space-time processing.

[1081] The DFE and MLSE techniques may reduce or possibly eliminate the degradation in performance when channel estimates are not as accurate. The DFE and MLSE techniques are described in further detail by S.L. Ariyavistakul *et al.* in a paper entitled "Optimum Space-Time Processors with Dispersive Interference: Unified Analysis and Required Filter Span," IEEE Trans. on Communication, Vol. 7, No. 7, July 1999, and incorporated herein by reference.

[1082] Adaptive transmitter processing based on the available CSI and successive cancellation receiver processing may also be advantageously employed for dispersive MIMO channels. The SNR for a recovered transmitted signal from the output of each space-time processing stage may comprise the CSI for that transmitted signal. This information may be fed back to the transmitter to aid in the selection of an appropriate coding and modulation scheme for the data stream associated with that transmitted signal.

Receiver Structure

[1083] FIG. 5 is a block diagram of a receiver system 150a capable of implementing various aspects and embodiments of the invention. Receiver system 150a implements

the successive cancellation receiver processing technique to receive and recover the transmitted signals. The transmitted signals from (up to) N_T transmit antennas are received by each of N_R antennas 152a through 152r and routed to a respective demodulator (DEMOD) 154 (which is also referred to as a front-end processor). For example, receive antenna 152a may receive a number of transmitted signals from a number of transmit antennas, and receive antenna 152r may similarly receive multiple transmitted signals. Each demodulator 154 conditions (e.g., filters and amplifies) a respective received signal, downconverts the conditioned signal to an intermediate frequency or baseband, and digitizes the downconverted signal to provide samples. Each demodulator 154 may further demodulate the samples with a received pilot to generate a stream of received modulation symbols, which is provided to RX MIMO/data processor 156.

[1084] If OFDM is employed for the data transmission, each demodulator 154 further performs processing complementary to that performed by modulator 122 shown in FIG. 3. In this case, each demodulator 154 includes an FFT processor (not shown) that generates transformed representations of the samples and provides a stream of modulation symbol vectors. Each vector includes N_L modulation symbols for N_L frequency subchannels and one vector is provided for each time slot. The modulation symbol vector streams from the FFT processors of all N_R demodulators are then provided to a demultiplexer (not shown in FIG. 5), which "channelizes" the modulation symbol vector stream from each FFT processor into a number of (up to N_L) modulation symbol streams. For the transmit processing scheme in which each frequency subchannel is independently processed (e.g., as shown in FIG. 3), the demultiplexer further provides each of (up to) N_L modulation symbol streams to a respective RX MIMO/data processor 156.

[1085] For a MIMO system utilizing OFDM, one RX MIMO/data processor 156 may be used to process the N_R modulation symbol streams from the N_R received antennas for each of the N_L frequency subchannels used for data transmission. And for a MIMO system not utilizing OFDM, one RX MIMO/data processor 156 may be used to process the N_R modulation symbol streams from the N_R received antennas.

[1086] In the embodiment shown in FIG. 5, RX MIMO/data processor 156 includes a number of successive (i.e., cascaded) receiver processing stages 510, one stage for each of the transmission channels used for data transmission. In one transmit

processing scheme, one data stream is transmitted on each transmission channel, and each data stream is independently processed (e.g., with its own encoding and modulation scheme) and transmitted from a respective transmit antenna. For this transmit processing scheme, the number of data streams is equal to the number of transmission channels, which is equal to the number of transmit antennas used for data transmission (which may be a subset of the available transmit antennas). For clarity, RX MIMO/data processor 156 is described for this transmit processing scheme.

[1087] Each receiver processing stage 510 (except for the last stage 510n) includes a channel MIMO/data processor 520 coupled to an interference canceller 530, and the last stage 510n includes only channel MIMO/data processor 520n. For the first receiver processing stage 510a, channel MIMO/data processor 520a receives and processes the N_R modulation symbol streams from demodulators 154a through 154r to provide a decoded data stream for the first transmission channel (or the first transmitted signal). And for each of the second through last stages 510b through 510n, channel MIMO/data processor 520 for that stage receives and processes the N_R modified symbol streams from the interference canceller in the preceding stage to derive a decoded data stream for the transmission channel being processed by that stage. Each channel MIMO/data processor 520 further provides CSI (e.g., the SNR) for the associated transmission channel.

[1088] For the first receiver processing stage 510a, interference canceller 530a receives the N_R modulation symbol streams from all N_R demodulators 154. And for each of the second through second-to-last stages, interference canceller 530 receives the N_R modified symbol streams from the interference canceller in the preceding stage. Each interference canceller 530 also receives the decoded data stream from channel MIMO/data processor 520 within the same stage, and performs the processing (e.g., encoding, interleaving, modulation, channel response, and so on) to derive N_R remodulated symbol streams that are estimates of the interference components of the received modulation symbol streams due to this decoded data stream. The remodulated symbol streams are then subtracted from the received modulation symbol streams to derive N_R modified symbol streams that include all but the subtracted (i.e., cancelled) interference components. The N_R modified symbol streams are then provided to the next stage.

[1089] In FIG. 5, a controller 540 is shown coupled to RX MIMO/data processor 156 and may be used to direct various steps in the successive cancellation receiver processing performed by processor 156.

[1090] FIG. 5 shows a receiver structure that may be used in a straightforward manner when each data stream is transmitted over a respective transmit antenna (i.e., one data stream corresponding to each transmitted signal). In this case, each receiver processing stage 510 may be operated to recover one of the transmitted signals and provide the decoded data stream corresponding to the recovered transmitted signal.

[1091] For some other transmit processing schemes, a data stream may be transmitted over multiple transmit antennas, frequency subchannels, and/or time intervals to provide spatial, frequency, and time diversity, respectively. For these schemes, the receiver processing initially derives a received modulation symbol stream for the transmitted signal on each transmit antenna of each frequency subchannel. Modulation symbols for multiple transmit antennas, frequency subchannels, and/or time intervals may be combined in a complementary manner as the demultiplexing performed at the transmitter system. The stream of combined modulation symbols is then processed to provide the associated decoded data stream.

Spatial Processing Techniques for Non-Dispersive Channels

[1092] As noted above, a number of linear spatial processing techniques may be used to process the signals received via a non-dispersive channel to recover each transmitted signal stream from interference caused by the other transmitted signal streams. These techniques include the CCMI, MMSE, and possibly other techniques. The linear spatial processing is performed within each channel MIMO/data processor 520 on the N_R input signals. For the first receiver processing stage 510a, the input signals are the N_R received signals from the N_R received antennas. And for each subsequent stage, the input signals are the N_R modified signals from the interference canceller from the preceding stage, as described above. For clarity, the CCMI and MMSE techniques are described for the first stage. However, the processing for each subsequent stage proceeds in similar manner with the proper substitution for the input signals. More specifically, at each subsequent stage the signals detected in the previous stage are assumed to be cancelled, so the dimensionality of the channel coefficient matrix is reduced at each stage as described below.

[1093] In a MIMO system with N_T transmit antennas and N_R receive antennas, the received signals at the output of the N_R receive antennas may be expressed as:

$$\underline{\mathbf{r}} = \underline{\mathbf{H}}\underline{\mathbf{x}} + \underline{\mathbf{n}} \ , \qquad \text{Eq (6)}$$

where $\underline{\mathbf{r}}$ is the received symbol vector (i.e., the N_Rx1 vector output from the MIMO channel, as derived from the receive antennas), $\underline{\mathbf{H}}$ is the channel coefficient matrix, $\underline{\mathbf{x}}$ is the transmitted symbol vector (i.e., the N_Tx1 vector input into the MIMO channel), and $\underline{\mathbf{n}}$ is an N_Rx1 vector representing noise plus interference. The received symbol vector $\underline{\mathbf{r}}$ includes N_R modulation symbols from N_R signals received via N_R receive antennas at a specific time slot. Similarly, the transmitted symbol vector $\underline{\mathbf{x}}$ includes N_T modulation symbols in N_T signals transmitted via N_T transmit antennas at a specific time slot.
The channel coefficient matrix $\underline{\mathbf{H}}$ can be further written as:

$$\underline{\mathbf{H}} = [\underline{\mathbf{h}}_1 \ \underline{\mathbf{h}}_2 \ \Lambda \ \ \underline{\mathbf{h}}_{N_T}] \qquad \text{Eq (6a)}$$

where the vectors $\underline{\mathbf{h}}_i$ contain the channel coefficients associated with the i-th transmit antenna. At each subsequent step in the successive cancellation process, the column vectors in equation (6a) associated with previously cancelled signals are removed. Assuming for simplicity that the transmit signals are cancelled in the same order that the associated channel coefficient vectors are listed in equation (6a), then at the k-th stage in the successive cancellation process, the channel coefficient matrix is:

$$\underline{\mathbf{H}} = [\underline{\mathbf{h}}_k \ \underline{\mathbf{h}}_{k+1} \ \Lambda \ \ \underline{\mathbf{h}}_{N_T}] \qquad \text{Eq (6b)}$$

CCMI Technique

[1094] For the CCMI spatial processing technique, the receiver system first performs a channel matched filter operation on the received symbol vector $\underline{\mathbf{r}}$. The matched-filtered output can be expressed as:

$$\underline{\mathbf{H}}^H \underline{\mathbf{r}} = \underline{\mathbf{H}}^H \underline{\mathbf{H}}\underline{\mathbf{x}} + \underline{\mathbf{H}}^H \underline{\mathbf{n}} \ , \qquad \text{Eq (7)}$$

where the superscript "^{H}" represents transpose and complex conjugate. A square matrix $\underline{\mathbf{R}}$ may be used to denote the product of the channel coefficient matrix $\underline{\mathbf{H}}$ with its conjugate-transpose $\underline{\mathbf{H}}^H$ (i.e., $\underline{\mathbf{R}} = \underline{\mathbf{H}}^H \underline{\mathbf{H}}$).

[1095] The channel coefficient matrix $\underline{\mathbf{H}}$ may be derived, for example, from pilot symbols transmitted along with the data. In order to perform "optimal" reception and to

estimate the SNR of the transmission channels, it is often convenient to insert some known symbols into the transmit data stream and to transmit the known symbols over one or more transmission channels. Such known symbols are also referred to as pilot symbols or pilot signals. Methods for estimating a single transmission channel based on a pilot signal and/or a data transmission may be found in a number of papers available in the art. One such channel estimation method is described by F. Ling in a paper entitled "Optimal Reception, Performance Bound, and Cutoff-Rate Analysis of References-Assisted Coherent CDMA Communications with Applications," IEEE Transaction On Communication, Oct. 1999. This or some other channel estimation method may be extended to matrix form to derive the channel coefficient matrix $\underline{\mathbf{H}}$, as is known in the art.

[1096] An estimate of the transmitted symbol vector, $\underline{\mathbf{x}}'$, may be obtained by multiplying the matched-filtered vector $\underline{\mathbf{H}}^H \underline{\mathbf{r}}$ with the inverse (or pseudo-inverse) of $\underline{\mathbf{R}}$, which can be expressed as:

$$\begin{aligned} \underline{\mathbf{x}}' &= \underline{\mathbf{R}}^{-1}\underline{\mathbf{H}}^H \underline{\mathbf{r}} \\ &= \underline{\mathbf{x}} + \underline{\mathbf{R}}^{-1}\underline{\mathbf{H}}^H \underline{\mathbf{n}} \qquad \text{Eq (8)} \\ &= \underline{\mathbf{x}} + \underline{\mathbf{n}}' \quad . \end{aligned}$$

From the above equation, it can be observed that the transmitted symbol vector $\underline{\mathbf{x}}$ may be recovered by matched filtering (i.e., multiplying with the matrix $\underline{\mathbf{H}}^H$) the received symbol vector $\underline{\mathbf{r}}$ and then multiplying the filtered result with the inverse square matrix $\underline{\mathbf{R}}^{-1}$.

[1097] For the CCMI technique, the SNR of the received symbol vector after processing (i.e., the i-th element of $\underline{\mathbf{x}}'$) can be expressed as:

$$SNR_i = \frac{\overline{|x'_i|^2}}{\sigma_{n'}^2} \quad . \qquad \text{Eq (9)}$$

If the variance of the i-th transmitted symbol $\overline{|x'_i|^2}$ is equal to one (1.0) on the average, the SNR of the receive symbol vector after processing may be expressed as:

$$SNR_i = \frac{1}{r_{ii}\sigma_n^2} \quad .$$

The noise variance may be normalized by scaling the i-th element of the received symbol vector by $1/\sqrt{\tilde{r}_{ii}}$.

[1098] If a modulation symbol stream was duplicated and transmitted over multiple transmit antennas, then these modulation symbols may be summed together to form combined modulation symbols. For example, if a data stream was transmitted from all antennas, then the modulation symbols corresponding to all N_T transmit antennas are summed, and the combined modulation symbol may be expressed as:

$$x'_{total} = \sum_{i=1}^{N_T} \frac{x'_i}{\tilde{r}_{ii}} \quad . \qquad \text{Eq (10)}$$

Alternatively, the transmitter may be operated to transmit one or more data streams on a number of transmission channels using the same coding and modulation scheme on some or all transmit antennas. In this case, only one SNR (e.g., an average SNR) may be needed for the transmission channels for which the common coding and modulation scheme is applied. For example, if the same coding and modulation scheme is applied on all transmit antennas, then the SNR of the combined modulation symbol, SNR_{total}, can be derived. This SNR_{total} would then have a maximal combined SNR that is equal to the sum of the SNR of the modulation symbols from the N_T transmit antennas. The combined SNR may be expressed as:

$$SNR_{total} = \sum_{i=1}^{N_T} SNR_i = \frac{1}{\sigma_n^2} \sum_{i=1}^{N_T} \frac{1}{\tilde{r}_{ii}} \quad . \qquad \text{Eq (11)}$$

[1099] FIG. 6A is a block diagram of an embodiment of a channel MIMO/data processor 520x, which is capable of implementing the CCMI technique described above. Channel MIMO/data processor 520x includes a processor 610x (which performs CCMI processing) coupled to a RX data processor 620.

[1100] Within processor 610x, the received modulation symbol vectors $\underline{\mathbf{r}}$ are filtered by a match filter 614, which pre-multiplies each vector $\underline{\mathbf{r}}$ with the conjugate-transpose channel coefficient matrix $\underline{\mathbf{H}}^H$, as shown above in equation (7). The channel coefficient matrix $\underline{\mathbf{H}}$ may be estimated based on pilot signals in a manner similar to that used for conventional pilot assisted single and multi-carrier systems, as is known in the art. The matrix $\underline{\mathbf{R}}$ is computed according to the equation $\underline{\mathbf{R}} = \underline{\mathbf{H}}^H \underline{\mathbf{H}}$, as shown above. The filtered vectors are further pre-multiplied by a multiplier 616 with the inverse square

matrix $\underline{\mathbf{R}}^{-1}$ to form an estimate $\underline{\mathbf{x}}'$ of the transmitted modulation symbol vector $\underline{\mathbf{x}}$, as shown above in equation (8).

[1101] For certain transmit processing schemes, the estimated modulation symbol streams corresponding to multiple transmit antennas used for the transmission of a data stream may be provided to a combiner 618, which combines redundant information across time, space, and frequency. The combined modulation symbols $\underline{\mathbf{x}}''$ are then provided to RX data processor 620. For some other transmit processing schemes, the estimated modulation symbols $\underline{\mathbf{x}}'$ may be provided directly (not shown in FIG. 6A) to RX data processor 620.

[1102] Processor 610x thus generates a number of independent symbol streams corresponding to the number of data streams transmitted from the transmitter system. Each symbol stream includes recovered modulation symbols that correspond to and are estimates of the modulation symbols after the symbol mapping at the transmitter system. The (recovered) symbol streams are then provided to RX data processor 620.

[1103] As noted above, each stage 510 within RX MIMO/data processor 156 recovers and decodes one of the transmitted signals (e.g., the transmitted signal with the best SNR) included in that stage's input signals. The estimation of the SNRs for the transmitted signals is performed by a CSI processor 626, and may be achieved based on equations (9) and (11) described above. CSI processor 626 then provides the CSI (e.g., SNR) for the transmitted signal that has been selected (e.g., the "best") for recovery and decoding, and further provides a control signal identifying the selected transmitted signal.

[1104] FIG. 7 is a block diagram of an embodiment of RX data processor 620. In this embodiment, a selector 710 within RX data processor 620 receives a number of symbol streams from a preceding linear spatial processor and extracts the symbol stream corresponding to the selected transmitted signal, as indicated by the control signal from CSI processor 626. In an alternative embodiment, RX data processor 620 is provided with the symbol stream corresponding to the selected transmitted signal and the stream extraction may be performed by combiner 618 based on the control signal from CSI processor 626. In any case, the extracted stream of modulation symbols is provided to a demodulation element 712.

[1105] For the transmitter embodiment shown in FIG. 2 in which the data stream for each transmission channel is independently coded and modulated based on the

channel's SNR, the recovered modulation symbols for the selected transmission channel are demodulated in accordance with a demodulation scheme (e.g., M-PSK, M-QAM) that is complementary to the modulation scheme used for the transmission channel. The demodulated data from demodulation element 712 is then de-interleaved by a de-interleaver 714 in a complementary manner to that performed by channel interleaver 214, and the de-interleaved data is further decoded by a decoder 716 in a complementary manner to that performed by encoder 212. For example, a Turbo decoder or a Viterbi decoder may be used for decoder 716 if Turbo or convolutional coding, respectively, is performed at the transmitter. The decoded data stream from decoder 716 represents an estimate of the transmitted data stream being recovered.

[1106] Referring back to FIG. 6A, the estimated modulation symbols $\underline{\mathbf{x}}'$ and/or the combined modulation symbols $\underline{\mathbf{x}}''$ are also provided to CSI processor 626, which estimates the SNR for each of the transmission channels. For example, CSI processor 626 may estimate the noise covariance matrix ϕ_{nn} based on the pilot signals received and then compute the SNR of the i-th transmission channel based on equation (9) or (11). The SNR can be estimated similar to conventional pilot assisted single and multi-carrier systems, as is known in the art. The SNR for all of the transmission channels may comprise the CSI that is reported back to the transmitter system for this transmission channel. CSI processor 626 further provides to RX data processor 620 or combiner 618 the control signal that identifies the selected transmission channel.

[1107] The estimated modulation symbols $\underline{\mathbf{x}}'$ are further provided to a channel estimator 622 and a matrix processor 624 that respectively estimates the channel coefficient matrix $\underline{\mathbf{H}}$ and derives the square matrix $\underline{\mathbf{R}}$. The estimated modulation symbols corresponding to pilot data and/or traffic data may be used for the estimation of the channel coefficient matrix $\underline{\mathbf{H}}$.

[1108] Referring back to FIG. 5, the input signals to the first stage 510a includes all transmitted signals, and the input signals to each subsequent stage includes one transmitted signal (i.e., one term) canceled by a preceding stage. Thus, channel MIMO/data processor 520a within the first stage 510a may be designed and operated to estimate the channel coefficient matrix $\underline{\mathbf{H}}$ and to provide this matrix to all subsequent stages.

[1109] The CSI information to be reported by receiver system 150 back to transmitter system 110 may comprise the SNRs for the transmission channels, as determined by the stages within RX MIMO/data processor 156.

MMSE Technique

[1110] For the MMSE spatial processing technique, the receiver system first performs a multiplication of the received symbol vector $\underline{\mathbf{r}}$ with a weighting coefficient matrix $\underline{\mathbf{M}}$ to derive an initial MMSE estimate $\hat{\underline{\mathbf{x}}}$ of the transmitted symbol vector $\underline{\mathbf{x}}$, which can be expressed as:

$$\begin{aligned}\hat{\underline{\mathbf{x}}} &= \underline{\mathbf{M}}\underline{\mathbf{r}} \\ &= \underline{\mathbf{H}}^H(\underline{\mathbf{H}}\underline{\mathbf{H}}^H + \underline{\phi}_{nn})^{-1}\underline{\mathbf{r}}\end{aligned} \qquad \text{Eq (12)}$$

where

$$\underline{\mathbf{M}} = \underline{\mathbf{H}}^H(\underline{\mathbf{H}}\underline{\mathbf{H}}^H + \underline{\phi}_{nn})^{-1} \qquad \text{Eq (13)}$$

The matrix $\underline{\mathbf{M}}$ is selected such that the mean square error of the error vector $\underline{\mathbf{e}}$ between the initial MMSE estimate $\hat{\underline{\mathbf{x}}}$ and the transmitted symbol vector $\underline{\mathbf{x}}$ (i.e., $\underline{\mathbf{e}} = \hat{\underline{\mathbf{x}}} - \underline{\mathbf{x}}$) is minimized.

[1111] To determine the SNR of the transmission channels for the MMSE technique, the signal component can first be determined based on the mean of $\hat{\underline{\mathbf{x}}}$ given $\underline{\mathbf{x}}$, averaged over the additive noise, which can be expressed as:

$$\begin{aligned}E[\hat{\underline{\mathbf{x}}} \mid \underline{\mathbf{x}}] &= E[\underline{\mathbf{M}}\underline{\mathbf{r}} \mid \underline{\mathbf{x}}] \\ &= \underline{\mathbf{H}}^H(\underline{\mathbf{H}}\underline{\mathbf{H}}^H + \underline{\phi}_{nn})^{-1}E[\underline{\mathbf{r}}] \\ &= \underline{\mathbf{H}}^H(\underline{\mathbf{H}}\underline{\mathbf{H}}^H + \underline{\phi}_{nn})^{-1}\underline{\mathbf{H}}\underline{\mathbf{x}} \\ &= \underline{\mathbf{V}}\underline{\mathbf{x}} \quad ,\end{aligned}$$

where the matrix $\underline{\mathbf{V}}$ can be expressed as:

$$\underline{\mathbf{V}} = \underline{\mathbf{H}}^H(\underline{\phi}_{nn} + \underline{\mathbf{H}}\underline{\mathbf{H}}^H)^{-1}\underline{\mathbf{H}} = \underline{\mathbf{H}}^H\underline{\phi}_{nn}^{-1}\underline{\mathbf{H}}(\underline{\mathbf{I}} + \underline{\mathbf{H}}^H\underline{\phi}_{nn}^{-1}\underline{\mathbf{H}})^{-1} \ .$$

[1112] The i-th element $\hat{x}_i$ of the initial MMSE estimate $\hat{\underline{\mathbf{x}}}$ can be expressed as:

$$\hat{x}_i = v_{i1}x_1 + \ldots + v_{ii}x_i + \ldots + v_{iN_R}x_{N_R} \ . \qquad \text{Eq (14)}$$

If all of the elements of $\hat{\underline{\mathbf{x}}}$ are uncorrelated and have zero mean, the expected value of the i-th element of $\hat{\underline{\mathbf{x}}}$ can be expressed as:

$$E[\hat{x}_i \mid \mathbf{x}] = v_{ii}x_i \ . \qquad \text{Eq (15)}$$

[1113] As shown in equation (15), $\hat{x}_i$ is a biased estimate of x_i and this bias can be removed to obtain improved performance. An unbiased estimate of x_i can be obtained by dividing $\hat{x}_i$ by v_{ii}. Thus, the unbiased minimum mean square error estimate of $\underline{\mathbf{x}}$, $\tilde{\underline{\mathbf{x}}}$, can be obtained by pre-multiplying the biased estimate $\hat{\underline{\mathbf{x}}}$ by a diagonal matrix $\underline{\mathbf{D}}_{\mathbf{V}}^{-1}$, as follows:

$$\tilde{\underline{\mathbf{x}}} = \underline{\mathbf{D}}_{\mathbf{V}}^{-1}\hat{\underline{\mathbf{x}}} \ , \qquad \text{Eq (16)}$$

where

$$\underline{\mathbf{D}}_{\mathbf{V}}^{-1} = diag(1/v_{11}, 1/v_{22}, \ldots, 1/v_{N_R N_R}) \ , \qquad \text{Eq (17)}$$

and v_{ii} are the diagonal elements of the matrix $\underline{\mathbf{V}}$.

[1114] To determine the noise plus interference, the error $\hat{\underline{\mathbf{e}}}$ between the unbiased estimate $\tilde{\underline{\mathbf{x}}}$ and the transmitted symbol vector $\underline{\mathbf{x}}$ can be expressed as:

$$\begin{aligned} \hat{\underline{\mathbf{e}}} &= \underline{\mathbf{x}} - \underline{\mathbf{D}}_{\mathbf{V}}^{-1}\hat{\underline{\mathbf{x}}} \\ &= \underline{\mathbf{x}} - \underline{\mathbf{D}}_{\mathbf{V}}^{-1}\underline{\mathbf{H}}^H(\underline{\mathbf{H}}\underline{\mathbf{H}}^H + \underline{\phi}_{nn})^{-1}\underline{\mathbf{r}} \end{aligned}$$

[1115] For the MMSE technique, the SNR of the received symbol vector after processing (i.e., the i-th element of $\tilde{\underline{\mathbf{x}}}$) can be expressed as:

$$SNR_i = \frac{E[\overline{|x_i|^2}]}{u_{ii}} \qquad \text{Eq (18)}$$

where u_{ii} is the variance of the i-th element of the error vector $\hat{\underline{\mathbf{e}}}$, and the matrix $\mathbf{U}$ can be expressed as:

$$\underline{\mathbf{U}} = \underline{\mathbf{I}} - \underline{\mathbf{D}}_{\mathbf{V}}^{-1}\underline{\mathbf{V}} - \underline{\mathbf{V}}\underline{\mathbf{D}}_{\mathbf{V}}^{-1} + \underline{\mathbf{D}}_{\mathbf{V}}^{-1}\underline{\mathbf{V}}\underline{\mathbf{D}}_{\mathbf{V}}^{-1} \ . \qquad \text{Eq (19)}$$

If the variance, $\overline{|x_i|^2}$, of the i-th transmitted symbol, x_i, is equal to one (1.0) on the average, and from equation (19) $u_{ii} = \frac{1}{v_{ii}} - 1$, then the SNR of the received symbol vector after processing may be expressed as:

$$SNR_i = \frac{v_{ii}}{1 - v_{ii}} . \qquad \text{Eq (20)}$$

The estimated modulation symbols, $\underline{\tilde{\mathbf{x}}}$, may similarly be combined to obtain combined modulation symbols, as described above for the CCMI technique.

[1116] FIG. 6B is a block diagram of an embodiment of a channel MIMO/data processor 520y, which is capable of implementing the MMSE technique described above. Channel MIMO/data processor 520y includes a processor 610y (which performs MMSE processing) coupled to RX data processor 620.

[1117] Within processor 610y, the received modulation symbol vectors $\underline{\mathbf{r}}$ are pre-multiplied with the matrix $\underline{\mathbf{M}}$ by a multiplier 634 to form an estimate $\underline{\hat{\mathbf{x}}}$ of the transmitted symbol vector $\underline{\mathbf{x}}$, as shown above in equation (8). Similar to the CCMI technique, the matrices $\underline{\mathbf{H}}$ and $\underline{\phi}_{nn}$ may be estimated based on the received pilot signals and/or data transmissions. The matrix $\underline{\mathbf{M}}$ is then computed according to equation (9). The estimate $\underline{\hat{\mathbf{x}}}$ is further pre-multiplied with the diagonal matrix $\underline{\mathbf{D}}_{\mathbf{V}}^{-1}$ by a multiplier 636 to form an unbiased estimate $\underline{\tilde{\mathbf{x}}}$ of the transmitted symbol vector $\underline{\mathbf{x}}$, as shown above in equation (12).

[1118] Again, for certain transmit processing schemes, a number of streams of estimated modulation symbols $\underline{\tilde{\mathbf{x}}}$ corresponding to a number of transmit antennas used for transmitting a data stream may be provided to a combiner 638, which combines redundant information across time, space, and frequency. The combined modulation symbols $\underline{\tilde{\mathbf{x}}}''$ are then provided to RX data processor 620. For some other transmit processing schemes, the estimated modulation symbols $\underline{\tilde{\mathbf{x}}}$ may be provided directly (not shown in FIG. 6B) to RX data processor 620. RX data processor 620 demodulates, de-interleaves, and decodes the modulation symbol stream corresponding to the data stream being recovered, as described above.

[1119] The estimated modulation symbols $\underline{\tilde{\mathbf{x}}}$ and/or the combined modulation symbols $\underline{\tilde{\mathbf{x}}}''$ are also provided to CSI processor 626, which estimates the SNR for each

of the transmitted signals. For example, CSI processor 626 may estimate the SNR of the i-th transmitted signal based on equation (18) or (20). The SNR for the selected transmitted signal may be reported back to the transmitter system. CSI processor 626 further provides to RX data processor 620 or combiner 618 the control signal that identifies the selected transmitted signal.

[1120] The estimated modulation symbols $\underline{\tilde{\mathbf{x}}}$ are further provided to an adaptive processor 642 that derives the matrix $\underline{\mathbf{M}}$ and the diagonal matrix $\underline{\mathbf{D}}_{\mathrm{V}}^{-1}$ based on equations (13) and (17), respectively.

Space-Time Processing Techniques for Time-Dispersive Channels

[1121] As noted above, a number of space-time processing techniques may be used to process the signals received via a time-dispersive channel. These techniques include the use of time domain channel equalization techniques such as MMSE-LE, DFE, MLSE, and possibly other techniques, in conjunction with the spatial processing techniques described above for a non-dispersive channel. The space-time processing is performed within each channel MIMO/data processor 520 on the N_R input signals.

MMSE-LE Technique

[1122] In the presence of time dispersion, the channel coefficient matrix $\underline{\mathbf{H}}$ takes on a delay dimension, and each element of the matrix $\underline{\mathbf{H}}$ behaves as a linear transfer function instead of a coefficient. In this case, the channel coefficient matrix $\underline{\mathbf{H}}$ can be written in the form of a channel transfer function matrix $\underline{\mathbf{H}}(\tau)$, which can be expressed as:

$$\underline{\mathbf{H}}(\tau) = \{h_{ij}(\tau)\} \quad \text{for } 1 \le i \le \mathrm{N_R}, \text{ and } 1 \le j \le \mathrm{N_T} \ . \qquad \text{Eq (21)}$$

where $h_{ij}(\tau)$ is the linear transfer function from the j-th transmit antenna to the i-th receive antenna. As a result of the linear transfer functions $h_{ij}(\tau)$, the received signal vector $\underline{\mathbf{r}}(t)$ is a convolution of the channel transfer function matrix $\underline{\mathbf{H}}(\tau)$ with the transmitted signal vector $\underline{\mathbf{x}}(t)$, which can be expressed as:

$$\underline{\mathbf{r}}(t) = \int \underline{\mathbf{H}}(\tau)\ \underline{\mathbf{x}}(t-\tau)d\tau \ . \qquad \text{Eq (22)}$$

[1123] As part of the demodulation function (performed by demodulators 154 in FIG. 5), the received signals are sampled to provide received samples. Without loss of generality, the time-dispersive channel and the received signals can be represented in a discrete-time representation in the following description. First, the channel transfer function vector $\underline{\mathbf{h}}_j(k)$ associated with the j-th transmit antenna at delay k can be expressed as:

$$\underline{\mathbf{h}}_j(k) = [h_{1j}(k)\ h_{2j}(k)\ \Lambda\ h_{N_R j}(k)]^T \qquad \text{for } 0 \leq k \leq L \ , \qquad \text{Eq (23)}$$

where $h_{ij}(k)$ is the k-th tap weight of the channel transfer function associated with the path between the j-th transmit antenna and the i-th receive antenna and L is the maximum extent (in sample intervals) of the channel time dispersion. Next, the $\mathrm{N_R} \times \mathrm{N_T}$ channel transfer function matrix at delay k can be expressed as:

$$\underline{\mathbf{H}}(k) = [\underline{\mathbf{h}}_1(k)\ \underline{\mathbf{h}}_2(k)\ \Lambda\ \underline{\mathbf{h}}_{\mathrm{N_T}}(k)] \qquad \text{for } 0 \leq k \leq L \qquad \text{Eq (24)}$$

[1124] The received signal vector $\underline{\mathbf{r}}(n)$ at sample time n can then be expressed as:

$$\underline{\mathbf{r}}(n) = \sum_{k=0}^{L} \underline{\mathbf{H}}(k)\underline{\mathbf{x}}(n-k) + \underline{\mathbf{n}}(n) = \underline{\underline{\mathbf{H}}}\underline{\underline{\mathbf{x}}}(n) + \underline{\mathbf{n}}(n) \ , \qquad \text{Eq (25)}$$

where $\underline{\underline{\mathbf{H}}}$ is an $\mathrm{N_R} \times (\mathrm{L}+1)\mathrm{N_T}$ block-structured matrix that represents the sampled channel matrix transfer function, $\underline{\mathbf{H}}(k)$, and can be represented as:

$$\underline{\underline{\mathbf{H}}} = [\underline{\mathbf{H}}(0)\ \underline{\mathbf{H}}(1)\ \Lambda\ \underline{\mathbf{H}}(L)] \ ,$$

and $\underline{\underline{\mathbf{x}}}(n)$ is a sequence of L+1 vectors of received samples captured for L+1 sample intervals, with each vector comprising $\mathrm{N_R}$ samples for the $\mathrm{N_R}$ received antennas, and can be represented as:

$$\underline{\underline{\mathbf{x}}}(n) = \begin{bmatrix} \underline{\mathbf{x}}(n) \\ \underline{\mathbf{x}}(n-1) \\ \mathrm{M} \\ \underline{\mathbf{x}}(n-L) \end{bmatrix} .$$

[1125] An MMSE linear space-time processor computes an estimate of the transmitted symbol vector, $\hat{\underline{\mathbf{x}}}(n)$, at time n by performing a convolution of the sequence

of received signal vectors $\underline{\mathbf{r}}(n)$ with the sequence of $2K+1$, $\mathrm{N_R} \times \mathrm{N_T}$ weight matrices $\underline{\mathbf{M}}(k)$, as follows:

$$\hat{\underline{\mathbf{x}}}(n) = \sum_{k=-K}^{K} \underline{\mathbf{M}}(k)\underline{\mathbf{r}}(n-k) = \underline{\underline{\mathbf{M}}}\underline{\underline{\mathbf{r}}}(n) \quad , \qquad \text{Eq (26)}$$

where $\underline{\underline{\mathbf{M}}} = [\underline{\mathbf{M}}(-K) \quad \Lambda \quad \underline{\mathbf{M}}(0) \quad \Lambda \quad \underline{\mathbf{M}}(K)]$, K is a parameter that determines the delay-extent of the equalizer filter, and

$$\underline{\underline{\mathbf{r}}}(n) = \begin{bmatrix} \underline{\mathbf{r}}(n+K) \\ \mathrm{M} \\ \underline{\mathbf{r}}(n) \\ \mathrm{M} \\ \underline{\mathbf{r}}(n-K) \end{bmatrix} .$$

The sequence of weight matrices $\underline{\mathbf{M}}(k)$ is selected to minimize the mean-square error, which can be expressed as:

$$\varepsilon = E\{\underline{\mathbf{e}}^H(k)\underline{\mathbf{e}}(k)\} , \qquad \text{Eq (27)}$$

where the error $\underline{\mathbf{e}}(k)$ can be expressed as:

$$\underline{\mathbf{e}}(k) = \hat{\underline{\mathbf{x}}}(k) - \underline{\mathbf{x}}(k) . \qquad \text{Eq (28)}$$

[1126] The MMSE solution can then be stated as the sequence of weight matrices $\underline{\mathbf{M}}(k)$ that satisfy the linear constraints:

$$\sum_{k=-K}^{K} \underline{\mathbf{M}}(k)\underline{\mathbf{R}}(k-\lambda) = \begin{cases} 0, & -K \le \lambda < -L \\ \underline{\mathbf{H}}^H(-\lambda), & -L \le \lambda \le 0 \\ 0, & 0 < \lambda \le K \end{cases} , \qquad \text{Eq (29)}$$

where $\underline{\mathbf{R}}(k)$ is a sequence of $\mathrm{N_R} \times \mathrm{N_R}$ space-time correlation matrices, which can be expressed as:

$$\underline{\mathbf{R}}(k) = E\{\underline{\mathbf{r}}(n-k)\underline{\mathbf{r}}^H(n)\} = \begin{cases} \displaystyle\sum_{m=\max(0,-k)}^{\min(L,L-k)} \underline{\mathbf{H}}(m)\underline{\mathbf{H}}^H(m+k) + \underline{\varphi}_{zz}(k), & -L \le k \le L \\ \underline{\varphi}_{zz}(k), & \text{otherwise} \end{cases} \qquad \text{Eq (30)}$$

where $\underline{\varphi}_{zz}(k)$ is the noise autocorrelation function, which can be expressed as:

$$\underline{\varphi}_{zz}(k) = E\{\underline{\mathbf{z}}(\lambda - k)\underline{\mathbf{z}}^H(\lambda)\} \quad . \tag{Eq (31)}$$

For white (temporally uncorrelated) noise, $\underline{\varphi}_{zz}(k) = \underline{\varphi}_{zz}\delta(k)$, where $\underline{\varphi}_{zz}$ in this case represents only the spatial correlation matrix. For spatially and temporally uncorrelated noise with equal power at each receive antenna, $\underline{\varphi}_{zz}(k) = \sigma^2\underline{\mathbf{I}}\delta(k)$.

[1127] Equation (29) can further be represented as:

$$\underline{\underline{\mathbf{MR}}} = \underline{\underline{\tilde{\mathbf{H}}}}^H \text{, or } \underline{\underline{\mathbf{M}}} = \underline{\underline{\tilde{\mathbf{H}}}}^H \underline{\underline{\mathbf{R}}}^{-1} \quad , \tag{Eq (32)}$$

where $\underline{\underline{\mathbf{R}}}$ is block-Toeplitz with block j,k given by $\underline{\mathbf{R}}(j-k)$ and

$$\underline{\underline{\tilde{\mathbf{H}}}} = \begin{bmatrix} \underline{\mathbf{0}}_{(K-L)N_R \times N_T} \\ \underline{\mathbf{H}}(L) \\ \underline{\mathbf{H}}(L-1) \\ \mathrm{M} \\ \underline{\mathbf{H}}(0) \\ \underline{\mathbf{0}}_{K,N_R \times N_T} \end{bmatrix} ,$$

where $\underline{\mathbf{0}}_{m \times n}$ is an $m \times n$ matrix of zeros.

[1128] As with the MMSE spatial processing described above, to determine the SNR associated with the symbol estimates, an unbiased minimum mean square error estimate is derived. First, for the MMSE-LE estimate derived above,

$$\begin{aligned} E\left[\,\underline{\hat{\mathbf{x}}}(n) \middle| \underline{\mathbf{x}}(n)\right] &= \underline{\underline{\mathbf{M}}} E\left[\,\underline{\mathbf{r}}(n) \middle| \underline{\mathbf{x}}(n)\right] \\ &= [\underline{\mathbf{M}}(-K)\underline{\underline{\mathbf{H}}}\underline{\mathbf{x}}(n+K) + \Lambda + \underline{\mathbf{M}}(0)\underline{\underline{\mathbf{H}}}\underline{\mathbf{x}}(n) + \Lambda + \underline{\mathbf{M}}(K)\underline{\underline{\mathbf{H}}}\underline{\mathbf{x}}(n-K)] \end{aligned} \; , \tag{Eq (33)}$$

where the expectation is taken over the noise. If it is assumed that the modulation symbols are uncorrelated in time and the expectation is taken over all intersymbol interference in the above (all transmitted signal components not transmitted at time n), then the expectation can be expressed as:

$$\begin{aligned} E\left[\hat{\underline{\mathbf{x}}}(n)\middle|\underline{\mathbf{x}}(n)\right] &= \underline{\underline{\mathbf{M}}}E\left[\underline{\mathbf{r}}(n)\middle|\underline{\mathbf{x}}(n)\right] \\ &= [\underline{\mathbf{M}}(0)\underline{\mathbf{H}}(0) + \underline{\mathbf{M}}(-1)\underline{\mathbf{H}}(1) + \Lambda + \underline{\mathbf{M}}(-L)\underline{\mathbf{H}}(L)]\underline{\mathbf{x}}(n) \\ &= \underline{\underline{\mathbf{M}}}\underline{\underline{\tilde{\mathbf{H}}}}\underline{\mathbf{x}}(n) \\ &= \underline{\mathbf{V}}\underline{\mathbf{x}}(n) \end{aligned} \qquad \text{Eq (34)}$$

where

$$\underline{\mathbf{V}} = \underline{\underline{\mathbf{M}}}\underline{\underline{\tilde{\mathbf{H}}}} = \underline{\underline{\tilde{\mathbf{H}}}}^H \underline{\underline{\mathbf{R}}}^{-1} \underline{\underline{\tilde{\mathbf{H}}}} \ .$$

[1129] Finally, after averaging over the interference from other spatial subchannels, the average value of the signal from the i-th transmit antenna at time n can be expressed as:

$$E[\hat{x}_i(n) \mid x_i(n)] = v_{ii} x_i(n) \ , \qquad \text{Eq (35)}$$

where v_{ii} is the i-th diagonal element of $\underline{\mathbf{V}}$ (v_{ii} is a scalar), and $\hat{x}_i(n)$ is the i-th element of the MMSE-LE estimate.

[1130] By defining

$$\underline{\mathbf{D}}_{\mathrm{V}}^{-1} = diag(1/v_{11}, 1/v_{22}, \Lambda, 1/v_{N_T N_T}) \ , \qquad \text{Eq (36)}$$

then the unbiased MMSE-LE estimate of the transmitted signal vector at time n can be expressed as:

$$\tilde{\underline{\mathbf{x}}}(n) = \underline{\mathbf{D}}_{\mathrm{V}}^{-1}\hat{\underline{\mathbf{x}}}(n) = \underline{\mathbf{D}}_{\mathrm{V}}^{-1}\underline{\underline{\mathbf{M}}}\underline{\mathbf{r}}(n) \ . \qquad \text{Eq (37)}$$

The error covariance matrix associated with the unbiased MMSE-LE can be expressed as:

$$\begin{aligned} \underline{\varphi}_{ee} = \underline{\mathbf{U}} = E\left\{\left[\underline{\mathbf{x}}(n) - \underline{\mathbf{D}}_{\mathrm{V}}^{-1}\underline{\underline{\mathbf{M}}}\underline{\mathbf{r}}(n)\right]\left[\underline{\mathbf{x}}(n) - \underline{\mathbf{r}}^H(n)\underline{\underline{\mathbf{M}}}^H\underline{\mathbf{D}}_{\mathrm{V}}^{-1}\right]\right\} \\ = \mathbf{I} - \underline{\mathbf{D}}_{\mathrm{V}}^{-1}\underline{\mathbf{V}} - \underline{\mathbf{V}}\underline{\mathbf{D}}_{\mathrm{V}}^{-1} + \underline{\mathbf{D}}_{\mathrm{V}}^{-1}\underline{\mathbf{V}}\underline{\mathbf{D}}_{\mathrm{V}}^{-1} \ . \end{aligned} \qquad \text{Eq (38)}$$

The SNR associated with the estimate of the symbol transmitted on the i-th transmit antenna can finally be expressed as:

$$SNR_i = \frac{1}{\mathbf{u}_{ii}} = \frac{\mathbf{v}_{ii}}{1 - \mathbf{v}_{ii}} \ . \qquad \text{Eq (39)}$$

[1131] The MMSE-LE technique can be implemented by channel MIMO/data processor 520y in FIG. 6B. In this case, multiplier 634 can be designed to perform the convolution of the sequence of received signal vectors $\underline{\mathbf{r}}(n)$ with the sequence of weight matrices $\underline{\mathbf{M}}(k)$, as shown in equation (26). Multiplier 636 can be designed to perform the pre-multiply of the estimate $\hat{\underline{\mathbf{x}}}$ with the diagonal matrix $\underline{\mathbf{D}}_V^{-1}$ to derive the unbiased MMSE-LE estimate $\tilde{\underline{\mathbf{x}}}$, as shown in equation (37). Adaptive processor 642 can be designed to derive the sequence of weight matrices $\underline{\mathbf{M}}(k)$ as shown in equation (32) and the diagonal matrix $\underline{\mathbf{D}}_V^{-1}$ as shown in equation (36). The subsequent processing may be achieved in similar manner as that described above for the MMSE technique. The SNR of the symbol stream transmitted from the j-th transmit antenna may be estimated based on equation (39) by CSI processor 626.

DFE Technique

[1132] FIG. 6C is a block diagram of an embodiment of a channel MIMO/data processor 520z, which is capable of implementing the DFE spatial-time equalization technique. Channel MIMO/data processor 520z includes a space-time processor 610z, which performs DFE processing, coupled to RX data processor 620.

[1133] For the DFE technique, the received modulation symbol vectors $\underline{\mathbf{r}}(n)$ are received and processed by a forward receive processor 654 to provide estimated modulation symbols for the data stream to be recovered. Forward receive processor 654 may implement the CCMI or MMSE technique described above or some other linear spatial equalization technique. A summer 656 then combines an estimated distortion components provided by a feedback processor 658 with the estimated modulation symbols to provide "equalized" modulation symbols having the distortion component removed. Initially, the estimated distortion components are zero and the equalized modulation symbols are simply the estimated modulation symbols. The equalized modulation symbols from summer 656 are then demodulated and decoded by RX data processor 620 to provide the decoded data stream.

[1134] The decoded data stream is then re-encoded and re-modulated by a channel data processor 210x to provide remodulated symbols, which are estimates of the modulation symbols at the transmitter. Channel data processor 210x performs the same processing (e.g., encoding, interleaving, and modulation) as that performed at the transmitter for the data stream, e.g., as shown in FIG. 2. The remodulated symbols from

channel data processor 210x are provided to feedback processor 658, which processes the symbols to derive the estimated distortion components. Feedback processor 658 may implement a linear spatial equalizer (e.g., a linear transversal equalizer).

[1135] The resulting estimate of the transmitted symbol vector at time n can be expressed as:

$$\hat{\underline{\mathbf{x}}}(n) = \sum_{k=-K_1}^{0} \underline{\mathbf{M}}_f(k)\underline{\mathbf{r}}(n-k) + \sum_{k=1}^{K_2} \underline{\mathbf{M}}_b(k)\tilde{\underline{\mathbf{x}}}(n-k) \quad , \qquad \text{Eq (40)}$$

where $\underline{\mathbf{r}}(n)$ is the vector of received modulation symbols, which is given above in equation (25), $\tilde{\underline{\mathbf{x}}}(n)$ is the vector of symbol decisions provided by the channel data processor 210x, $\underline{\mathbf{M}}_f(k), -K_1 \leq k \leq 0$ is the sequence of (K_1+1)-$(\mathrm{N_T} \times \mathrm{N_R})$ feed-forward coefficient matrices used by forward receive processor 654, and $\underline{\mathbf{M}}_b(k), 1 \leq k \leq K_2$ is the sequence of K_2-$(\mathrm{N_T} \times \mathrm{N_R})$ feed-back coefficient matrices used by feedback processor 658. Equation (40) can also be expressed as:

$$\hat{\underline{\mathbf{x}}}(n) = \underline{\underline{\mathbf{M}}}_f \underline{\underline{\mathbf{r}}}(n) + \underline{\underline{\mathbf{M}}}_b \tilde{\underline{\underline{\mathbf{x}}}}(n) \quad , \qquad \text{Eq (41)}$$

where $\underline{\underline{\mathbf{M}}}_f = [\underline{\mathbf{M}}(-K_1)\ \underline{\mathbf{M}}(-K_1+1)\ \Lambda\ \underline{\mathbf{M}}(0)]$, $\underline{\underline{\mathbf{M}}}_b = [\underline{\mathbf{M}}(1)\ \underline{\mathbf{M}}(2)\ \Lambda\ \underline{\mathbf{M}}(K_2)]$,

$$\tilde{\underline{\underline{\mathbf{x}}}}(n) = \begin{bmatrix} \tilde{\underline{\mathbf{x}}}(n-1) \\ \tilde{\underline{\mathbf{x}}}(n-2) \\ \mathrm{M} \\ \tilde{\underline{\mathbf{x}}}(n-K_2) \end{bmatrix}, \text{ and } \underline{\underline{\mathbf{r}}}(n) = \begin{bmatrix} \underline{\mathbf{r}}(n+K_1) \\ \underline{\mathbf{r}}(n+K_1-1) \\ \mathrm{M} \\ \underline{\mathbf{r}}(n) \end{bmatrix}.$$

[1136] If the MMSE criterion is used to find the coefficient matrices, then the solutions for $\underline{\underline{\mathbf{M}}}_f$ and $\underline{\underline{\mathbf{M}}}_b$ that minimize the mean square error $\varepsilon = E\{\underline{\mathbf{e}}^H(k)\underline{\mathbf{e}}(k)\}$ can be used, where the error $\underline{\mathbf{e}}(k)$ is expressed as:

$$\underline{\mathbf{e}}(k) = \hat{\underline{\mathbf{x}}}(k) - \underline{\mathbf{x}}(k) \quad .$$

The MMSE solution for the feed-forward filter can then be expressed as:

$$\underline{\underline{\mathbf{M}}}_f = \tilde{\underline{\underline{\mathbf{H}}}}^H \tilde{\underline{\underline{\mathbf{R}}}}^{-1} \qquad \text{Eq (42)}$$

where

$$\underline{\underline{\tilde{\mathbf{H}}}} = \begin{bmatrix} \underline{\mathbf{0}}_{(K_1-L)N_R \times N_T} \\ \underline{\mathbf{H}}(L) \\ \underline{\mathbf{H}}(L-1) \\ \mathrm{M} \\ \underline{\mathbf{H}}(0) \end{bmatrix},$$

and $\underline{\underline{\tilde{\mathbf{R}}}}$ is a $(K_1+1)\mathrm{N_R} \times (K_1+1)\mathrm{N_R}$ matrix made up of $\mathrm{N_R} \times \mathrm{N_R}$ blocks. The *(i,j)*-th block in $\underline{\underline{\tilde{\mathbf{R}}}}$ is given by:

$$\underline{\tilde{\mathbf{R}}}(i,j) = \sum_{m=0}^{K_1-i+1} \underline{\mathbf{H}}(m)\underline{\mathbf{H}}^H(m+i-j) + \sigma^2 \underline{\mathbf{I}}\delta(i-j) \quad . \qquad \text{Eq (43)}$$

[1137] The MMSE solution for the feed-back filter is:

$$\underline{\mathbf{M}}_b(k) = -\sum_{j=-K_1}^{0} \underline{\mathbf{M}}_f(j)\underline{\mathbf{H}}(k-j),\ 1 \le k \le K_2 \quad . \qquad \text{Eq (44)}$$

As in the MMSE-LE described above, the unbiased estimate is first determined by finding the conditional mean value of the transmitted symbol vector:

$$E[\hat{\underline{\mathbf{x}}}(n) \mid \underline{\mathbf{x}}(n)] = \underline{\underline{\mathbf{M}}}_f \underline{\underline{\tilde{\mathbf{H}}}}\underline{\mathbf{x}}(n) = \underline{\mathbf{V}}_{\text{dfe}}\underline{\mathbf{x}}(n) \ , \qquad \text{Eq (45)}$$

where $\underline{\mathbf{V}}_{\text{dfe}} = \underline{\underline{\mathbf{M}}}_f \underline{\underline{\tilde{\mathbf{H}}}} = \underline{\underline{\tilde{\mathbf{H}}}}^H \underline{\underline{\tilde{\mathbf{R}}}}^{-1} \underline{\underline{\tilde{\mathbf{H}}}}$. Next, the mean value of the *i*-th element of $\hat{\underline{\mathbf{x}}}(n)$, $\hat{x}_i(n)$, is expressed as:

$$E[\hat{x}_i(n) \mid x_i(n)] = v_{\text{dfe},ii} x_i(n) \ ,$$

where $v_{\text{dfe},ii}$ is the *i*-th diagonal element of $\underline{\mathbf{V}}_{\text{dfe}}$. To form the unbiased estimate, similar to that described above, the diagonal matrix whose elements are the inverse of the diagonal elements of $\underline{\mathbf{V}}_{\text{dfe}}$ is first defined as:

$$\underline{\mathbf{D}}_{\text{Vdfe}}^{-1} = diag(v_{\text{dfe},11}, v_{\text{dfe},22}, \Lambda\ , v_{\text{dfe},N_T N_T}) \quad . \qquad \text{Eq (46)}$$

Then the unbiased estimate is expressed as:

$$\hat{\hat{\underline{\mathbf{x}}}}(n) = \underline{\mathbf{D}}_{\text{Vdfe}}^{-1}\hat{\underline{\mathbf{x}}}(n) = \underline{\mathbf{D}}_{\text{Vdfe}}^{-1}\underline{\underline{\mathbf{M}}}_f \underline{\underline{\mathbf{r}}}(n) + \underline{\mathbf{D}}_{\text{Vdfe}}^{-1}\underline{\underline{\mathbf{M}}}_b \underline{\underline{\tilde{\mathbf{x}}}}(n) \quad . \qquad \text{Eq (47)}$$

The resulting error covariance matrix is given by:

$$\underline{\varphi}_{ee} = \underline{\mathbf{U}}_{\text{dfe}} = E\left\{\left[\underline{\mathbf{x}}(n) - \underline{\mathbf{D}}_{\text{Vdfe}}^{-1}\left(\underline{\underline{\mathbf{M}}}_{f}\underline{\mathbf{r}}(n) + \underline{\underline{\mathbf{M}}}_{b}\underline{\tilde{\mathbf{x}}}(n)\right)\right]\left[\underline{\mathbf{x}}^{H}(n) - \left(\underline{\mathbf{r}}^{H}(n)\underline{\underline{\mathbf{M}}}_{f}^{H} + \underline{\tilde{\mathbf{x}}}^{H}(n)\underline{\underline{\mathbf{M}}}_{b}^{H}\right)\underline{\mathbf{D}}_{\text{Vdfe}}^{-1}\right]\right\}$$

$$= \mathbf{I} - \underline{\mathbf{D}}_{\text{Vdfe}}^{-1}\underline{\mathbf{V}}_{\text{dfe}} - \underline{\mathbf{V}}_{\text{dfe}}\underline{\mathbf{D}}_{\text{Vdfe}}^{-1} + \underline{\mathbf{D}}_{\text{Vdfe}}^{-1}\underline{\mathbf{V}}_{\text{dfe}}\underline{\mathbf{D}}_{\text{Vdfe}}^{-1}$$

Eq (48)

The SNR associated with the estimate of the symbol transmitted on the i-th transmit antenna can then be expressed as:

$$SNR_i = \frac{1}{u_{\text{dfe},ii}} = \frac{v_{\text{dfe},ii}}{1 - v_{\text{dfe},ii}} \quad . \qquad \text{Eq (49)}$$

[1138] With the DFE technique, the decoded data stream is used to derive an estimate of the distortion generated by the already decoded information bits. If the data stream is decoded without errors (or with minimal errors), then the distortion component can be accurately estimated and the inter-symbol interference contributed by the already decoded information bits may be effectively canceled out. The processing performed by forward receive processor 654 and feedback processor 658 are typically adjusted simultaneously to minimize the mean square error (MSE) of the inter-symbol interference in the equalized modulation symbols. DFE processing is described in further detail in the aforementioned paper by Ariyavistakul *et al.*

Interference Cancellation

[1139] FIG. 8 is a block diagram of an interference canceller 530x, which is one embodiment of interference canceller 530 in FIG. 5. Within interference canceller 530x, the decoded data stream from the channel MIMO/data processor 520 within the same stage is re-encoded, interleaved, and re-modulated by a channel data processor 210y to provide remodulated symbols, which are estimates of the modulation symbols at the transmitter prior to the MIMO processing and channel distortion. Channel data processor 210y performs the same processing (e.g., encoding, interleaving, and modulation) as that performed at the transmitter for the data stream. The remodulated symbols are then provided to a channel simulator 810, which processes the symbols with the estimated channel response to provide estimates of the interference due the decoded data stream.

[1140] For a non-dispersive channel, channel simulator 810 multiples the remodulated symbol stream associated with the i-th transmit antenna with the vector $\underline{\hat{\mathbf{h}}}_i$,

which is an estimate of the channel response between the i-th transmit antenna for which the data stream is being recovered and each of the N_R receive antennas. The vector $\underline{\hat{\mathbf{h}}}_i$ may be expressed as:

$$\underline{\hat{\mathbf{h}}}_i = \begin{bmatrix} \hat{h}_{i,1} \\ \hat{h}_{i,2} \\ \vdots \\ \hat{h}_{i,N_R} \end{bmatrix} , \qquad \text{Eq (50)}$$

and is one column of an estimated channel response matrix $\underline{\hat{\mathbf{H}}}$ that can be expressed as:

$$\underline{\hat{\mathbf{H}}} = \begin{bmatrix} \hat{h}_{1,1} & \hat{h}_{2,1} & \cdots & \hat{h}_{N_T,1} \\ \hat{h}_{1,2} & \hat{h}_{2,2} & \cdots & \hat{h}_{N_T,2} \\ \vdots & \vdots & \ddots & \vdots \\ \hat{h}_{1,N_R} & \hat{h}_{2,N_R} & \cdots & \hat{h}_{N_T,N_R} \end{bmatrix} . \qquad \text{Eq (51)}$$

The matrix $\underline{\hat{\mathbf{H}}}$ may be provided by the channel MIMO/data processor 520 within the same stage.

[1141] If the remodulated symbol stream corresponding to the i-th transmit antenna is expressed as $\breve{x}_i$, then the estimated interference component $\underline{\hat{\mathbf{i}}}^i$ due to the recovered transmitted signal may be expressed as:

$$\underline{\hat{\mathbf{i}}}^i = \begin{bmatrix} \hat{h}_{i,1} \cdot \breve{x}_i \\ \hat{h}_{i,2} \cdot \breve{x}_i \\ \vdots \\ \hat{h}_{i,N_R} \cdot \breve{x}_i \end{bmatrix} . \qquad \text{Eq (52)}$$

[1142] The N_R elements in the interference vector $\underline{\hat{\mathbf{i}}}^i$ correspond to the component of the received signal at each of the N_R receive antennas due to symbol stream transmitted on the i-th transmit antenna. Each element of the vector represents an estimated component due to the decoded data stream in the corresponding received modulation symbol stream. These components are interference to the remaining (not yet detected) transmitted signals in the N_R received modulation symbol streams (i.e., the vector $\underline{\mathbf{r}}^k$), and are subtracted (i.e., canceled) from the received signal vector $\underline{\mathbf{r}}^k$ by a

summer 812 to provide a modified vector $\underline{\mathbf{r}}^{k+1}$ having the components from the decoded data stream removed. This cancellation can be expressed as shown above in equation (5). The modified vector $\underline{\mathbf{r}}^{k+1}$ is provided as the input vector to the next receiver processing stage, as shown in FIG. 5.

[1143] For a dispersive channel, the vector $\hat{\underline{\mathbf{h}}}_i$ is replaced with an estimate of the channel transfer function vector defined in equation (23), $\hat{\underline{\mathbf{h}}}_i(k)$, $0 \le k \le L$. Then the estimated interference vector at time n, $\hat{\underline{\mathbf{i}}}^{\,i}(n)$, may be expressed as:

$$\hat{\underline{\mathbf{i}}}^{\,i}(n) = \begin{bmatrix} \sum_{k=0}^{L} \hat{h}_{i1}(k)\breve{x}_i(n-k) \\ \sum_{k=0}^{L} \hat{h}_{i2}(k)\breve{x}_i(n-k) \\ \vdots \\ \sum_{k=0}^{L} \hat{h}_{iN_R}(k)\breve{x}_i(n-k) \end{bmatrix}, \qquad \text{Eq (53)}$$

where $\breve{x}_i(n)$ is the remodulated symbol for time n. Equation (54) effectively convolves the remodulated symbols with the channel response estimates for each transmit-receive antenna pair.

[1144] For simplicity, the receiver architecture shown in FIG. 5 provides the (received or modified) modulation symbol streams to each receiver processing stage 510, and these streams have the interference components due to previously decoded data streams removed (i.e., canceled). In the embodiment shown in FIG. 5, each stage removes the interference components due to the data stream decoded by that stage. In some other design, the received modulation symbol streams may be provided to all stages, and each stage may perform the cancellation of interference components from all previously decoded data streams (which may be provided from preceding stages). The interference cancellation may also be skipped for one or more stages (e.g., if the SNR for the data stream is high). Various modifications to the receiver architecture shown in FIG. 5 may be made and are within the scope of the invention.

Deriving and Reporting CSI

[1145] For simplicity, various aspects and embodiments of the invention have been described above wherein the CSI comprises SNR. In general, the CSI may comprise

any type of information that is indicative of the characteristics of the communication link. Various types of information may be provided as CSI, some examples of which are described below.

[1146] In one embodiment, the CSI comprises signal-to-noise-plus-interference ratio (SNR), which is derived as the ratio of the signal power over the noise plus interference power. The SNR is typically estimated and provided for each transmission channel used for data transmission (e.g., each transmit data stream), although an aggregate SNR may also be provided for a number of transmission channels. The SNR estimate may be quantized to a value having a particular number of bits. In one embodiment, the SNR estimate is mapped to an SNR index, e.g., using a look-up table.

[1147] In another embodiment, the CSI comprises signal power and interference plus noise power. These two components may be separately derived and provided for each transmission channel used for data transmission.

[1148] In yet another embodiment, the CSI comprises signal power, interference power, and noise power. These three components may be derived and provided for each transmission channel used for data transmission.

[1149] In yet another embodiment, the CSI comprises signal-to-noise ratio plus a list of interference powers for each observable interference term. This information may be derived and provided for each transmission channel used for data transmission.

[1150] In yet another embodiment, the CSI comprises signal components in a matrix form (e.g., $N_T \times N_R$ complex entries for all transmit-receive antenna pairs) and the noise plus interference components in matrix form (e.g., $N_T \times N_R$ complex entries). The transmitter unit may then properly combine the signal components and the noise plus interference components for the appropriate transmit-receive antenna pairs to derive the quality for each transmission channel used for data transmission (e.g., the post-processed SNR for each transmitted data stream, as received at the receiver unit).

[1151] In yet another embodiment, the CSI comprises a data rate indicator for the transmit data stream. The quality of a transmission channel to be used for data transmission may be determined initially (e.g., based on the SNR estimated for the transmission channel) and a data rate corresponding to the determined channel quality may then be identified (e.g., based on a look-up table). The identified data rate is indicative of the maximum data rate that may be transmitted on the transmission channel for the required level of performance. The data rate is then mapped to and

represented by a data rate indicator (DRI), which can be efficiently coded. For example, if (up to) seven possible data rates are supported by the transmitter unit for each transmit antenna, then a 3-bit value may be used to represent the DRI where, e.g., a zero can indicate a data rate of zero (i.e., don't use the transmit antenna) and 1 through 7 can be used to indicate seven different data rates. In a typical implementation, the quality measurements (e.g., SNR estimates) are mapped directly to the DRI based on, e.g., a look-up table.

[1152] In yet another embodiment, the CSI comprises an indication of the particular processing scheme to be used at the transmitter unit for each transmit data stream. In this embodiment, the indicator may identify the particular coding scheme and the particular modulation scheme to be used for the transmit data stream such that the desired level of performance is achieved.

[1153] In yet another embodiment, the CSI comprises a differential indicator for a particular measure of quality for a transmission channel. Initially, the SNR or DRI or some other quality measurement for the transmission channel is determined and reported as a reference measurement value. Thereafter, monitoring of the quality of the transmission channel continues, and the difference between the last reported measurement and the current measurement is determined. The difference may then be quantized to one or more bits, and the quantized difference is mapped to and represented by the differential indicator, which is then reported. The differential indicator may indicate to increase or decrease the last reported measurement by a particular step size (or to maintain the last reported measurement). For example, the differential indicator may indicate that (1) the observed SNR for a particular transmission channel has increased or decreased by a particular step size, or (2) the data rate should be adjusted by a particular amount, or some other change. The reference measurement may be transmitted periodically to ensure that errors in the differential indicators and/or erroneous reception of these indicators do not accumulate.

[1154] Other forms of CSI may also be used and are within the scope of the invention. In general, the CSI includes sufficient information in whatever form that may be used to adjust the processing at the transmitter such that the desired level of performance is achieved for the transmitted data streams.

[1155] The CSI may be derived based on the signals transmitted from the transmitter unit and received at the receiver unit. In an embodiment, the CSI is derived

based on a pilot reference included in the transmitted signals. Alternatively or additionally, the CSI may be derived based on the data included in the transmitted signals.

[1156] In yet another embodiment, the CSI comprises one or more signals transmitted on the reverse link from the receiver unit to the transmitter unit. In some systems, a degree of correlation may exist between the forward and reverse links (e.g. time division duplexed (TDD) systems where the uplink and downlink share the same band in a time division multiplexed manner). In these systems, the quality of the forward link may be estimated (to a requisite degree of accuracy) based on the quality of the reverse link, which may be estimated based on signals (e.g., pilot signals) transmitted from the receiver unit. The pilot signals would then represent a means for which the transmitter could estimate the CSI as observed by the receiver unit.

[1157] The signal quality may be estimated at the receiver unit based on various techniques. Some of these techniques are described in the following patents, which are assigned to the assignee of the present application and incorporated herein by reference:

- U.S Patent No. 5,799,005, entitled "SYSTEM AND METHOD FOR DETERMINING RECEIVED PILOT POWER AND PATH LOSS IN A CDMA COMMUNICATION SYSTEM," issued August 25, 1998,
- U.S. Patent No. 5,903,554, entitled "METHOD AND APPARATUS FOR MEASURING LINK QUALITY IN A SPREAD SPECTRUM COMMUNICATION SYSTEM," issued May 11, 1999,
- U.S. Patent Nos. 5,056,109, and 5,265,119, both entitled "METHOD AND APPARATUS FOR CONTROLLING TRANSMISSION POWER IN A CDMA CELLULAR MOBILE TELEPHONE SYSTEM," respectively issued October 8, 1991 and November 23, 1993, and
- U.S Patent No. 6,097,972, entitled "METHOD AND APPARATUS FOR PROCESSING POWER CONTROL SIGNALS IN CDMA MOBILE TELEPHONE SYSTEM," issued August 1, 2000.

[1158] Various types of information for CSI and various CSI reporting mechanisms are also described in U.S Patent Application Serial No. 08/963,386, entitled "METHOD AND APPARATUS FOR HIGH RATE PACKET DATA TRANSMISSION," filed November 3, 1997, assigned to the assignee of the present application, and in

"TIE/EIA/IS-856 cdma2000 High Rate Packet Data Air Interface Specification", both of which are incorporated herein by reference.

[1159] The CSI may be reported back to the transmitter using various CSI transmission schemes. For example, the CSI may be sent in full, differentially, or a combination thereof. In one embodiment, CSI is reported periodically, and differential updates are sent based on the prior transmitted CSI. In another embodiment, the CSI is sent only when there is a change (e.g., if the change exceeds a particular threshold), which may lower the effective rate of the feedback channel. As an example, the SNRs may be sent back (e.g., differentially) only when they change. For an OFDM system (with or without MIMO), correlation in the frequency domain may be exploited to permit reduction in the amount of CSI to be fed back. As an example for an OFDM system, if the SNR corresponding to a particular spatial subchannel for N_M frequency subchannels is the same, the SNR and the first and last frequency subchannels for which this condition is true may be reported. Other compression and feedback channel error recovery techniques to reduce the amount of data to be fed back for CSI may also be used and are within the scope of the invention.

[1160] Referring back to FIG. 1, the CSI (e.g., the channel SNR) determined by RX MIMO processor 156 is provided to a TX data processor 162, which processes the CSI and provides processed data to one or more modulators 154. Modulators 154 further condition the processed data and transmit the CSI back to transmitter system 110 via a reverse channel.

[1161] At system 110, the transmitted feedback signal is received by antennas 124, demodulated by demodulators 122, and provided to a RX data processor 132. RX data processor 132 performs processing complementary to that performed by TX data processor 162 and recovers the reported CSI, which is then provided to, and used to adjust the processing by, TX data processor 114 and TX MIMO processor 120.

[1162] Transmitter system 110 may adjust (i.e., adapt) its processing based on the CSI (e.g., SNR information) from receiver system 150. For example, the coding for each transmission channel may be adjusted such that the information bit rate matches the transmission capability supported by the channel SNR. Additionally, the modulation scheme for the transmission channel may be selected based on the channel SNR. Other processing (e.g., interleaving) may also be adjusted and are within the scope of the invention. The adjustment of the processing for each transmission channel

based on the determined SNR for the channel allows the MIMO system to achieve high performance (i.e., high throughput or bit rate for a particular level of performance). The adaptive processing can be applied to a single-carrier MIMO system or a multi-carrier based MIMO system (e.g., a MIMO system utilizing OFDM).

[1163] The adjustment in the coding and/or the selection of the modulation scheme at the transmitter system may be achieved based on numerous techniques, one of which is described in the aforementioned U.S Patent Application Serial No. 09/776,975.

MIMO System Operating Schemes

[1164] Various operating schemes may be implemented for a MIMO system that employs adaptive transmitter processing (which is dependent on the available CSI) and successive cancellation receiver processing techniques described herein. Some of these operating schemes are described in further detail below.

[1165] In one operating scheme, the coding and modulation scheme for each transmission channel is selected based on the channel's transmission capability, as determined by the channel's SNR. This scheme can provide improved performance when used in combination with the successive cancellation receiver processing technique, as described in further detail below. When there is a large disparity between the worst-case and best-case transmission channels (i.e., transmit-receive antenna pairings), the coding may be selected to introduce sufficient redundancy to allow the receiver system to recover the original data stream. For example, the worst transmit antenna may have associated with it a poor SNR at the receiver output. The forward error correction (FEC) code is then selected to be powerful enough to allow the symbols transmitted from the worst-case transmit antenna to be correctly received at the receiver system. In practice, improved error correction capability comes at the price of increased redundancy, which implies a sacrifice in overall throughput. Thus, there is a tradeoff in terms of reduced throughput for increased redundancy using FEC coding.

[1166] When the transmitter is provided with the SNR per recovered transmitted signal, a different coding and/or modulation scheme may be used for each transmitted signal. For example, a specific coding and modulation scheme may be selected for each transmitted signal based on its SNR so that the error rates associated with the transmitted signals are approximately equal. In this way, throughput is not dictated by the SNR of the worst-case transmitted signal.

[1167] As an example, consider a 4 x 4 MIMO system with 4 transmit and 4 receive antennas and employing the successive cancellation receiver processing technique described herein. For this example, the SNR for the four transmitted signals are 5 dB, 8.5 dB, 13 dB, and 17.5 dB. If the same coding and modulation scheme is used for all four transmitted signal, the selected scheme would be dictated by the transmitted signal having 5 dB SNR. Using the information given in Table 1, each transmit antenna would employ a coding rate of 3/4 and QPSK modulation, giving a total modulation efficiency of 6 information bits/symbol, or 1.5 information bits/symbol/transmitted signal.

[1168] With CSI available, the transmitter may select the following coding and modulation schemes for the four transmitted signals, as shown in Table 2.

Table 2

SNR (dB)	Coding Rate	Modulation Symbol	# of Information Bits/Symbol
5	3/4	QPSK	1.5
8.5	5/8	16-QAM	2.5
13	7/12	64-QAM	3.5
17.5	5/6	64-QAM	5

By adjusting the coding and modulation scheme at the transmitter based on the available CSI, the effective modulation efficiency achieved is more than doubled to 12.5 bits/symbol versus 6 bits/symbol without CSI. The decoded error rate for each of the transmitted signals will be approximately equal since the coding and modulation scheme was selected to achieve this level of performance.

[1169] With adaptive processing at the transmitter system based on the available CSI, the successive cancellation receiver processing technique may be altered to take advantage of the fact that the bit error rates for the transmitted signals are approximately equal. If the coding and modulation scheme used on each transmitted signal provides an equivalent decoded error rate, then the ranking procedure (i.e., highest to lower SNR) may be omitted from the receiver processing, which may simplify the processing. In practical implementation, there may be slight differences in the decoded error rates for the transmitted signals. In this case, the SNR for the transmitted signals (after the linear or non-linear processing) may be ranked and the best post-processed SNR selected for detection (i.e., demodulation and decoding) first, as described above.

[1170] With CSI available at the transmitter, throughput is no longer dictated by the worst-case transmitted signal since the coding and modulation schemes are selected to provide a particular level of performance (e.g., a particular BER) on each transmission channel based on the channel's SNR. Since FEC coding is applied to each transmission channel independently, the minimum amount of redundancy required to meet the target level of performance is used, and throughput is maximized. The performance achievable with adaptive transmitter processing based on CSI (e.g., SNR) and successive cancellation receiver processing rivals that of a full-CSI processing scheme (whereby full characterization is available for each transmit-receive antenna pair) under certain operating conditions, as described in further detail below.

[1171] In another operating scheme, the transmitter it not provided with the SNR achieved for each transmission channel, but may be provided with a single value indicative of the average SNR for all transmission channels, or possibly some information indicating which transmit antennas to be used for data transmission. In this scheme, the transmitter may employ the same coding and modulation scheme on all transmit antennas used for data transmission, which may be a subset of the N_T available transmit antennas. When the same coding and modulation scheme is used on all transmit antennas, performance may be compromised. This is because the overall performance of the successive cancellation receiver processing technique is dependent on the ability to decode each transmitted signal error free. This correct detection is important to effectively cancel the interference due to the recovered transmitted signal.

[1172] By using same coding and modulation scheme for all transmitted signals, the recovered transmitted signal with the worst SNR will have the highest decoded error rate. This ultimately limits the performance of the MIMO system since the coding and modulation scheme is selected so that the error rate associated with the worst-case transmitted signal meets the overall error rate requirements. To improve efficiency, additional receive antennas may be used to provide improved error rate performance on the first recovered transmitted signal. By employ more receive antennas than transmit antennas, the error rate performance of the first recovered transmitted signal has a diversity order of (N_R-N_T+1) and reliability is increased.

[1173] In yet another operating scheme, the transmitted data streams are "cycled" across all available transmit antennas. This scheme improves the SNR statistics for each of the recovered transmitted signals since the transmitted data is not subjected to

the worst-case transmission channel, but instead is subjected to all transmission channels. The decoder associated with a specific data stream is effectively presented with "soft decisions" that are representative of the average across all possible pairs of transmit-receive antennas. This operating scheme is described in further detail in European Patent Application Serial No. 99302692.1, entitled "WIRELESS COMMUNICATIONS SYSTEM HAVING A SPACE-TIME ARCHITECTURE EMPLOYING MULTI-ELEMENT ANTENNAS AT BOTH THE TRANSMITTER AND RECEIVER," and incorporated herein by reference.

[1174] The successive cancellation receiver processing technique allows a MIMO system to utilize the additional dimensionalities created by the use of multiple transmit and receive antennas, which is a main advantage for employing MIMO. Depending on the characteristics of the MIMO channel, a linear spatial equalization technique (e.g., CCMI or MMSE) or a space-time equalization technique (e.g., MMSE-LE, DFE, or MLSE) may be used to process the received signals. The successive cancellation receiver processing technique, when used in combination with the adaptive transmitter processing based on the available CSI, may allow the same number of modulation symbols to be transmitted for each time slot as for a MIMO system utilizing full CSI.

[1175] Other linear and non-linear receiver processing techniques may also be used in conjunction with the successive cancellation receiver processing technique and the adaptive transmitter processing technique, and this is within the scope of the invention. Analogously, FIGS. 6A through 6C represent embodiments of three receiver processing technique capable of processing a MIMO transmission and determining the characteristics of the transmission channels (i.e., the SNR). Other receiver designs based on the techniques presented herein and other receiver processing techniques can be contemplated and are within the scope of the invention.

[1176] The linear and non-linear receiver processing techniques (e.g., CCMI, MMSE, MMSE-LE, DFE, MLSE, and other techniques) may also be used in a straightforward manner without adaptive processing at the transmitter when only the overall received signal SNR or the attainable overall throughput estimated based on such SNR is feed back. In one implementation, a modulation format is determined based on the received SNR estimate or the estimated throughput, and the same modulation format is used for all transmission channels. This method may reduce the

overall system throughput but may also greatly reduce the amount of information sent back over the reverse link.

System Performance

[1177] Improvement in system performance may be realized with the use of the successive cancellation receiver processing technique and the adaptive transmitter processing technique based on the available CSI. The system throughput with CSI feedback can be computed and compared against the throughput with full CSI feedback. The system throughput can be defined as:

$$C = \sum_{i=1}^{N_C} \log_2(1+\gamma_i) \ , \qquad \text{Eq (54)}$$

where γ_i is the SNR of each received modulation symbol. The SNR for some of the receiver processing techniques are summarized above.

[1178] FIG. 9A shows the improvement in SNR for a 4x4 MIMO channel configuration using the successive cancellation receiver processing technique. The results are obtained from a computer simulation. In the simulation, the following assumptions are made: (1) independent Rayleigh fading channels between receiver-transmit antenna pairs (i.e., no array correlation), (2) total interference cancellation (i.e., no decision errors are made in the decoding process and accurate channel estimates are available at the receiver). In practical implementation, channel estimates are not totally accurate, and a back-off factor may be used in the modulation scheme selected for each transmitted data stream. In addition, some decision errors are likely to occur in the detection of each transmitted data stream. This probability can be reduced if independently transmitted data streams are individually coded, which would then allow the receiver to decode the data streams independently, which may then reduce the probability of decision errors. In this case, the decoded data is re-encoded to construct the interference estimate used in the successive interference cancellation.

[1179] As shown in FIG. 9A, the first recovered transmitted signal has the poorest SNR distribution. Each subsequent recovered transmitted signal has improved SNR distributions, with the final recovered transmitted signal (i.e., the fourth one in this example) having the best overall SNR distribution. The distribution of the average SNR formed by summing the SNRs for the individual transmitted signals and dividing by four is also shown. The SNR distribution achieved without successive spatial

equalization and interference cancellation is given by the SNR distribution for the first recovered transmitted signal. In comparing the SNR distribution for the first recovered transmitted signal to the average SNR distribution, it can be seen that the spatial equalization and interference cancellation technique improves the effective SNR at the receiver.

[1180] *FIG. 9B shows the average throughput for a number of receive processing* techniques, including (1) the linear spatial equalization technique (without interference cancellation), (2) the spatial equalization and interference cancellation technique, and (3) the full-CSI technique. For each of these schemes, the transmitter is provided with either full or partial CSI for all transmitted signals, and the data for each transmitted signal is encoded and modulated based on the SNR. For the plots shown in FIG. 9B, the CCMI and MMSE techniques are used for the linear spatial equalization technique.

[1181] FIG. 9B shows the theoretical capacity (plot 920) achieved when using full-CSI processing based on the decomposition of the MIMO channel into eigenmodes. FIG. 9B further shows that the throughputs for both the CCMI technique (plot 924) and MMSE technique (plot 922) with partial-CSI but without interference cancellation have lower throughput than the capacity bound (plot 920).

[1182] Since capacity is proportional to SNR, as shown in equation (20), and SNR improves with the use of successive interference cancellation, capacity on average improves using the spatial equalization and interference cancellation technique. Using spatial equalization (with CCMI) and interference cancellation technique and partial-CSI, the throughput (plot 926) is improved over the spatial equalization only schemes (plots 922 and 924), with performance improving more as SNR increases. Using spatial equalization (with MMSE) and interference cancellation technique and partial-CSI, the throughput (plot 928) is identical to the capacity bound (plot 920), which represents remarkable system performance. Plot 920 assumes perfect channel estimates and no decision errors. The throughput estimates shown in FIG. 9B for the successive spatial equalization and interference cancellation technique with partial-CSI processing may degrade under practical implementations due to imperfect interference cancellation and detection errors.

[1183] FIG. 9C shows the average throughput for the successive space-time equalization (with MMSE-LE) and interference cancellation technique with adaptive transmitter processing based on CSI for a 4x4 MIMO system. The plots are obtained by

averaging over a large number of static realizations of a dispersive channel model (i.e., VehA). FIG. 9C shows the capacity bound (plot 930) and the performance of the MMSE-LE technique with interference cancellation (plot 934) and without successive interference cancellation (plot 932). The throughput performance for the MMSE-LE without successive interference cancellation technique (plot 932) degrades at higher SNR values. The throughput performance for the MMSE-LE with successive interference cancellation technique (plot 934) is close to channel capacity, which represents a high level of performance.

[1184] The elements of the transmitter and receiver systems may be implemented with one or more digital signal processors (DSP), application specific integrated circuits (ASIC), processors, microprocessors, controllers, microcontrollers, field programmable gate arrays (FPGA), programmable logic devices, other electronic units, or any combination thereof. Some of the functions and processing described herein may also be implemented with software executed on a processor.

[1185] Certain aspects of the invention may be implemented with a combination of software and hardware. For example, computations for the symbol estimates for the linear spatial equalization, the space-time equalization, and the derivation of the channel SNR may be performed based on program codes executed on a processor (controllers 540 in FIG. 5).

[1186] For clarity, the receiver architecture shown in FIG. 5 includes a number of receiving processing stages, one stage for each data stream to be decoded. In some implementations, these multiple stages may be implemented with a single hardware unit or a single software module that is re-executed for each stage. In this manner, the hardware or software may be time shared to simplify the receiver design.

[1187] Headings are included herein for reference and to aid in the locating certain sections. These heading are not intended to limit the scope of the concepts described therein under, and these concepts may have applicability in other sections throughout the entire specification.

[1188] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not

intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[1189] **WHAT IS CLAIMED IS:**

CLAIMS

1. A method for processing data at a receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

processing a plurality of input signals having included therein one or more symbol streams corresponding to one or more data streams to provide a decoded data stream for one of the one or more symbol streams;

deriving a plurality of modified signals based on the input signals and having components due to the decoded data stream approximately removed;

performing the processing and selectively performing the deriving for each of one or more iterations, one iteration for each data stream to be decoded, and wherein the input signals for each iteration subsequent to a first iteration are the modified signals from a preceding iteration; and

determining channel state information (CSI) indicative of characteristics of a MIMO channel used for transmitting the data steams, wherein the data streams are adaptively processed at a transmitter unit based in part on the CSI.

2. The method of claim 1, wherein the deriving is omitted for a last iteration.

3. The method of claim 1, wherein the processing includes

processing the input signals in accordance with a particular receive processing scheme to provide the one or more symbol streams, and

processing a selected one of the one or more symbol streams to provide the decoded data stream.

4. The method of claim 3, further comprising:

for each iteration,

estimating a quality of each of one or more unprocessed symbol streams included in the input signals; and

selecting one unprocessed symbol stream for processing based on estimated qualities for the one or more unprocessed symbol streams.

5. The method of claim 4, wherein the quality of each unprocessed symbol stream is estimated based on a signal-to-noise-plus-interference-ratio (SNR).

6. The method of claim 4, wherein the unprocessed symbol stream having the best estimated quality is selected for processing.

7. The method of claim 3, wherein the receive processing scheme performs linear spatial processing on the input signals.

8. The method of claim 7, wherein the receive processing scheme implements a channel correlation matrix inversion (CCMI) technique.

9. The method of claim 7, wherein the receive processing scheme implements a minimum mean square error (MMSE) technique

10. The method of claim 7, wherein the receive processing scheme implements a full-CSI processing technique

11. The method of claim 3, wherein the receive processing scheme performs space-time processing on the input signals.

12. The method of claim 11, wherein the receive processing scheme implements a minimum mean-square error linear space-time equalizer (MMSE-LE).

13. The method of claim 11, wherein the receive processing scheme implements a decision feedback space-time equalizer (DFE).

14. The method of claim 11, wherein the receive processing scheme implements a maximum-likelihood sequence estimator (MLSE).

15. The method of claim 1, wherein the deriving includes

generating a remodulated symbol stream based on the decoded data stream;

forming a plurality of interference signals based on the remodulated symbol stream; and

removing the interference signals from the input signals to derive the modified signals that serve as input signals for a succeeding iteration.

16. The method of claim 15, wherein the interference signals are formed based on a channel coefficient matrix $\underline{\mathbf{H}}$ indicative of characteristics of the MIMO channel.

17. The method of claim 1, further comprising:

transmitting the CSI from the receiver unit to the transmitter unit.

18. The method of claim 1, wherein the CSI comprises signal-to-noise-plus-interference-ratio (SNR) estimates for each of one or more transmission channels that compose the MIMO channel.

19. The method of claim 1, wherein the CSI comprises characterizations for one or more transmission channels that compose the MIMO channel.

20. The method of claim 1, wherein the CSI comprises an indication of a particular data rate supported by each of one or more transmission channels used for data transmission.

21. The method of claim 1, wherein the CSI comprises an indication of a particular processing scheme to be used for each of one or more transmission channels.

22. The method of claim 1, wherein the CSI comprises signal measurements and noise plus interference measurements for one or more transmission channels.

23. The method of claim 1, wherein the CSI comprises signal measurements, noise measurements, and interference measurements for one or more transmission channels.

24. The method of claim 1, wherein the CSI comprises signal to noise ratio and interference measurements for one or more transmission channels.

25. The method of claim 1, wherein the CSI comprises signal components and noise plus interference components for one or more transmission channels.

26. The method of claim 1, wherein the CSI comprises indications of changes in the characteristics of one or more transmission channels.

27. The method of claim 1, wherein the CSI is determined at the receiver unit and reported to the transmitter unit.

28. The method of claim 1, wherein the CSI is determined at the transmitter unit based on one or more signals transmitted by the receiver unit.

29. The method of claim 1, wherein each data stream is coded at the transmitter unit in accordance with a coding scheme selected based on the CSI for the transmission channel used to transmit the data stream.

30. The method of claim 29, wherein each data stream is independently coded in accordance with a coding scheme selected based on the CSI for the transmission channel used to transmit the data stream.

31. The method of claim 29, wherein each data stream is further modulated in accordance with a modulation scheme selected based on the CSI for the transmission channel used to transmit the data stream.

32. The method of claim 31, wherein the coding and modulation schemes are selected at the transmitter unit based on the CSI.

33. The method of claim 32, wherein the coding and modulation schemes are indicated by the CSI.

34. The method of claim 3, wherein the processing of the selected symbol stream includes

demodulating the symbol stream to provide demodulated symbols, and

decoding the demodulated symbols to provide the decoded data stream.

35. The method of claim 34, wherein the processing of the selected symbol stream further includes

deinterleaving the demodulated symbols, wherein the decoding is performed on the deinterleaved symbols to provide the decoded data stream.

36. The method of claim 1, wherein the MIMO system implements orthogonal frequency division modulation (OFDM).

37. The method of claim 36, wherein the processing at the receiver unit is independently performed for each of a plurality of frequency subchannels.

38. A method for processing data at a receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

receiving a plurality of signals via a plurality of received antennas;

processing the received signals in accordance with a particular receive processing scheme to provide a plurality of symbol streams corresponding to a plurality of transmitted data streams;

processing a selected one of the symbol streams to provide a decoded data stream;

forming a plurality of interference signals based on the decoded data stream;

deriving a plurality of modified signals based on the received signals and the interference signals;

performing the processing of the received signals and the selected symbol stream and selectively performing the forming and deriving for one or more iterations, one iteration each transmitted data stream to be decoded, wherein a first iteration is performed on the received signals and each subsequent iteration is performed on the modified signals from a preceding iteration; and

determining channel state information (CSI) indicative of characteristics of a MIMO channel used for transmitting the data steams, wherein the data streams are adaptively processed at a transmitter unit based in part on the CSI.

39. A method for communicating data from a transmitter unit to a receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

at the receiver unit,

receiving a plurality of signals via a plurality of receive antennas, wherein each received signal comprises a combination of one or more signals transmitted from the transmitter unit,

processing the received signals in accordance with a successive cancellation receiver processing technique to provide a plurality of decoded data streams transmitted from the transmitter unit,

determining channel state information (CSI) indicative of characteristics of a MIMO channel used to transmit the data steams, and

transmitting the CSI back to the transmitter unit; and

at the transmitter unit,

adaptively processing each data stream prior to transmission over the MIMO channel in accordance with the received CSI.

40. The method of claim 39, wherein the successive cancellation receiver processing scheme performs a plurality of iterations to provide the decoded data streams, one iteration for each decoded data stream.

41. The method of claim 40, wherein each iteration includes

processing a plurality of input signals in accordance with a particular linear or non-linear processing scheme to provide one or more symbol streams,

processing a selected one of the one or more symbol streams to provide a decoded data stream, and

deriving a plurality of modified signals based on the input signals and having components due to the decoded data stream approximately removed, wherein the input signals for a first iteration are the received signals and the input signals for each subsequent iteration are the modified signals from a preceding iteration.

42. The method of claim 39, wherein the CSI comprises a signal-to-noise-plus-interference-ratio (SNR) for each of one or more transmission channels that compose the MIMO channel.

43. The method of claim 39, wherein the CSI comprises an indication of a particular data rate supported by each of one or more transmission channels that compose the MIMO channel.

44. The method of claim 39, wherein the CSI comprises an indication of a particular processing scheme to be used for each of one or more transmission channels that compose the MIMO channel.

45. The method of claim 39, wherein the adaptive processing at the transmitter unit includes

encoding a data stream in accordance with a particular coding scheme selected based on the CSI associated with the data stream.

46. The method of claim 45, wherein the adaptive processing at the transmitter unit further includes

modulating the encoded data stream in accordance with a particular modulation scheme selected based on the CSI associated with the data stream.

47. A multiple-input multiple-output (MIMO) communication system, comprising:

a receiver unit comprising

a plurality of front-end processors configured to process a plurality of received signals to provide a plurality of symbol streams,

at least one receive processor coupled to the front-end processors and configured to process the symbol streams in accordance with a successive cancellation receiver processing scheme to provide a plurality of decoded data streams, and to further derive channel state information (CSI) indicative of characteristics of a MIMO channel used to transmit the data streams, and

a transmit data processor operatively coupled to the receive processor and configured to process the CSI for transmission back to the transmitter unit; and

a transmitter unit comprising

at least one demodulator configured to receive and process one or more signals from the receiver unit to recover the transmitted CSI, and

a transmit data processor configured to adaptively process data for transmission to the receiver unit based on the recovered CSI.

48. A receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

a plurality of front-end processors configured to process a plurality of received signals to provide a plurality of received symbol streams;

at least one receive processor coupled to the front-end processors and configured to process the received symbol streams to provide a plurality of decoded data streams, each receive processor including a plurality of processing stages, each stage configured to process input symbol streams to provide a respective decoded data stream and channel state information (CSI) associated with the decoded data stream, and to selectively provide modified symbol streams for a succeeding stage, wherein the input symbol streams for each stage are either the received symbol streams or the modified symbol streams from a preceding stage; and

a transmit processor configured to receive and process the CSI associated with the decoded data streams for transmission from the receiver unit, wherein the data streams are adaptively processed prior to transmission based in part on the CSI.

49. The receiver unit of claim 48, wherein each processing stage except a last stage includes

a channel processor configured to process the input symbol streams to provide a decoded data stream, and

an interference canceller configured to derive the modified symbol streams based on the decoded data stream and the input symbol streams.

50. The receiver unit of claim 49, wherein each channel processor includes

an input processor configured to process the input symbol streams to provide a recovered symbol stream, and

a data processor configured to process the recovered symbol stream to provide the decoded data stream.

51. The receiver unit of claim 50, wherein each input processor includes

a first processor configured to process the input symbol streams in accordance with a linear or non-linear receive processing scheme to provide the recovered symbol stream, and

a channel quality estimator configured to estimate a quality of the recovered symbol stream.

52. The receiver unit of claim 51, wherein the estimated quality comprises a signal-to-noise-plus-interference-ratio (SNR).

53. The receiver unit of claim 51, wherein the channel quality estimator is further configured to provide an indication of a data rate supported for the recovered symbol stream based on the quality estimate.

54. The receiver unit of claim 51, wherein the channel quality estimator is further configured to provide an indication of a particular processing scheme to be used at a transmitter unit for the recovered symbol stream based on the quality estimate.

55. The receiver unit of claim 51, wherein the estimated quality comprises an error signal indicative of detected noise plus interference level at the output of the receiver unit.

56. The receiver unit of claim 51, wherein the first processor performs linear spatial processing on the input symbol streams.

57. The receiver unit of claim 51, wherein the first processor performs space-time processing on the input symbol streams.

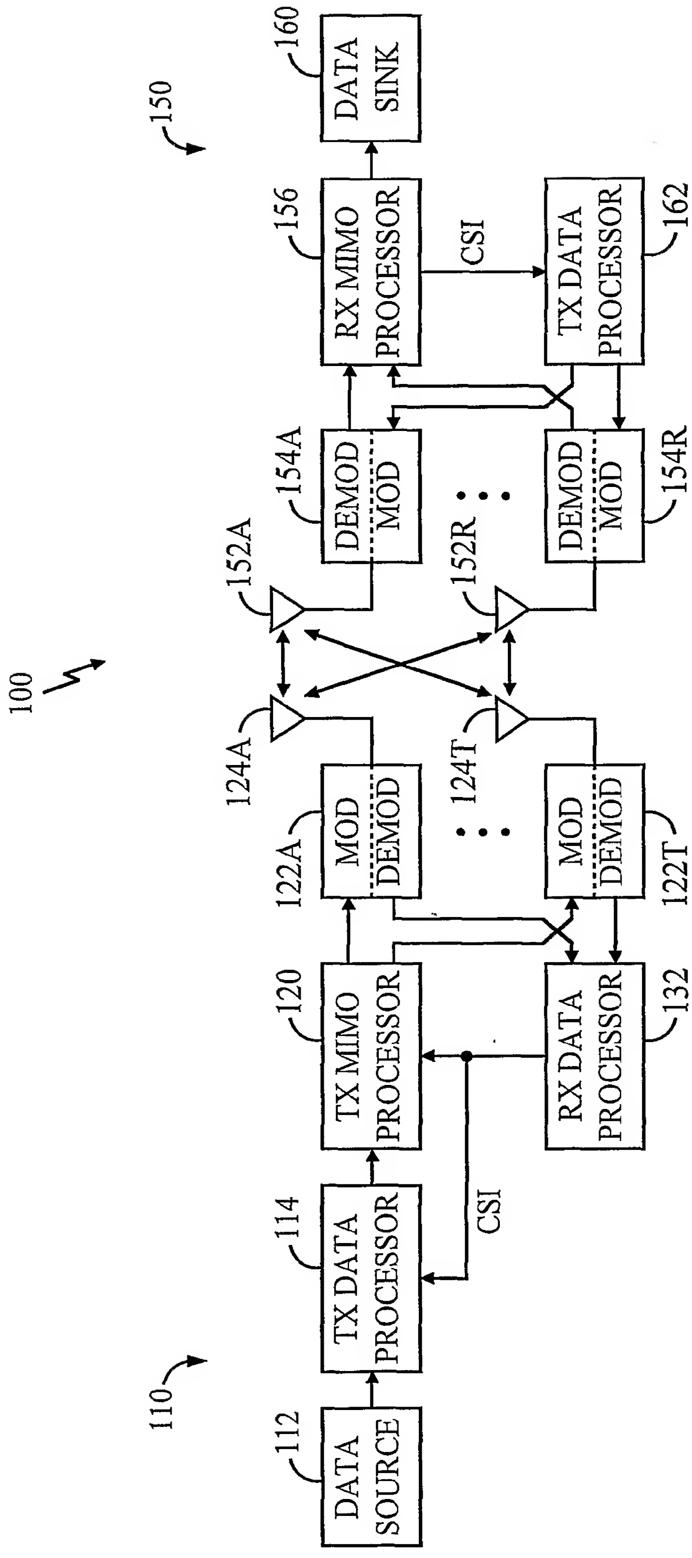
100
110
112
DATA SOURCE
114
TX DATA PROCESSOR
CSI
120
TX MIMO PROCESSOR
132
RX DATA PROCESSOR
122A
MOD
DEMOD
122T
MOD
DEMOD
124A
124T
152A
152R
154A
DEMOD
MOD
154R
DEMOD
MOD
150
156
RX MIMO PROCESSOR
CSI
162
TX DATA PROCESSOR
160
DATA SINK

FIG. 1

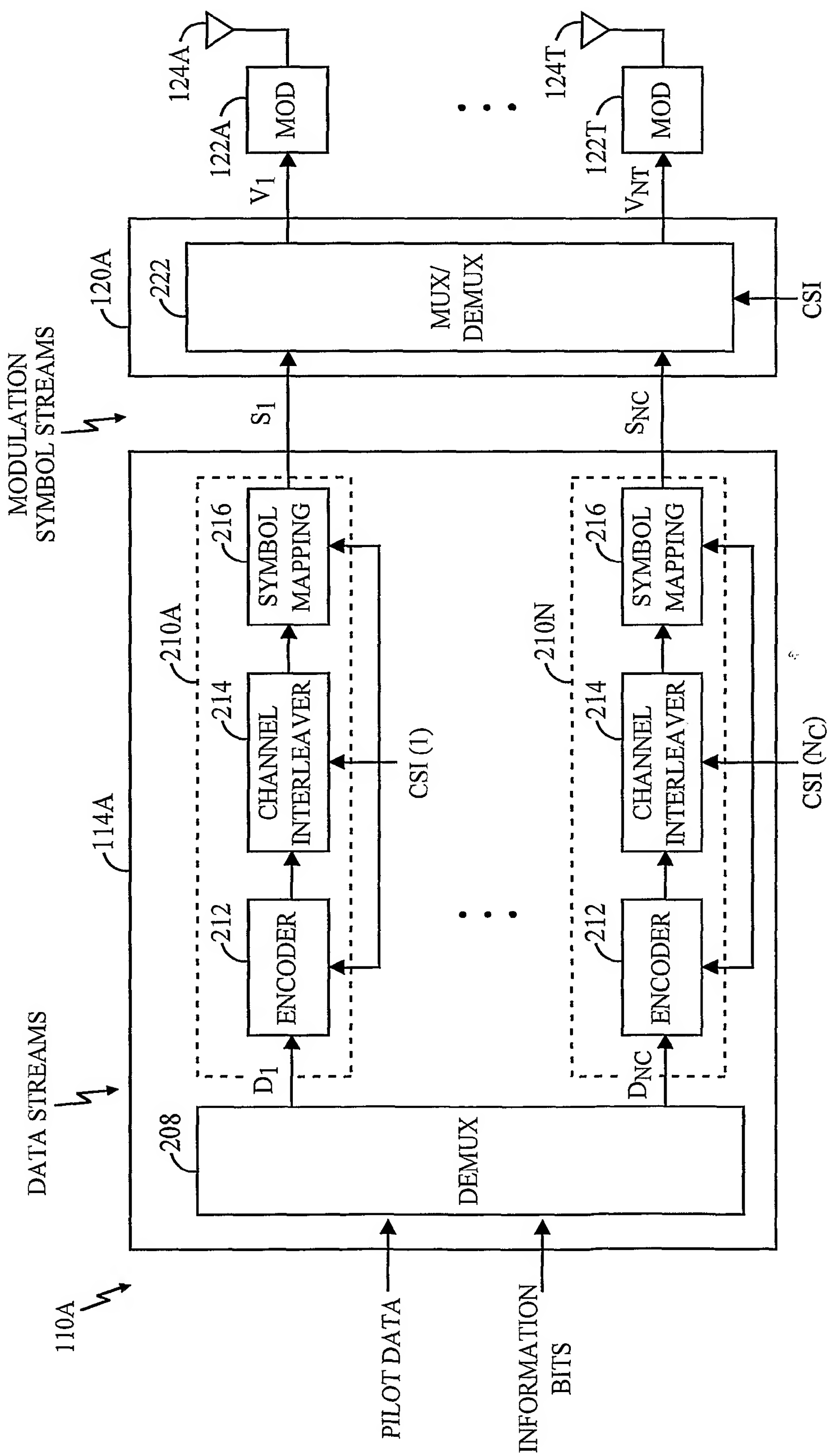
110A
PILOT DATA
INFORMATION BITS
DATA STREAMS
114A
208
DEMUX
D1
DNC
210A
210N
212
ENCODER
214
CHANNEL INTERLEAVER
216
SYMBOL MAPPING
CSI (1)
CSI (NC)
MODULATION SYMBOL STREAMS
S1
SNC
120A
222
MUX/ DEMUX
CSI
V1
VNT
122A
122T
MOD
124A
124T

FIG. 2

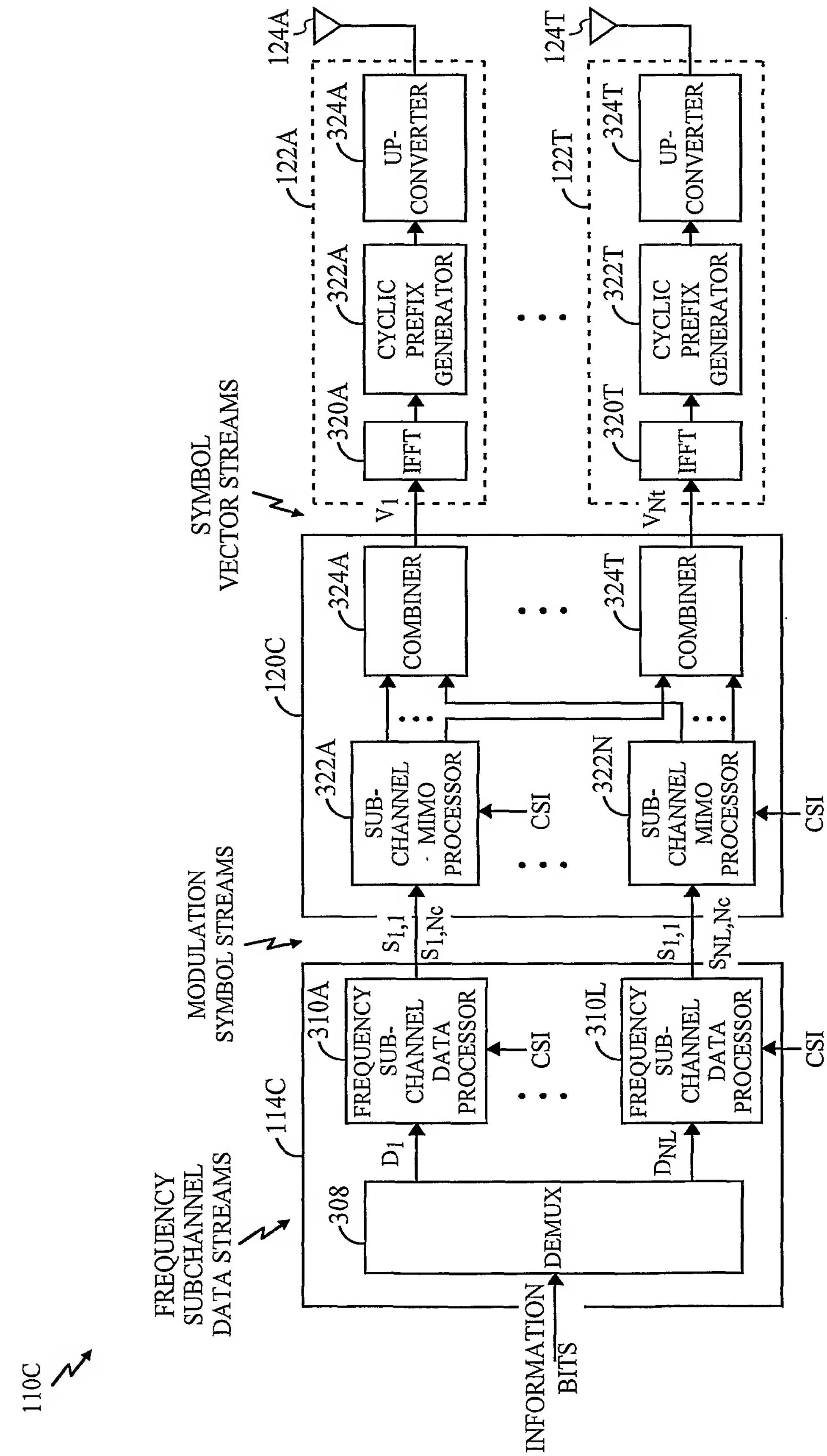
110C
FREQUENCY SUBCHANNEL DATA STREAMS
INFORMATION BITS
114C
308
DEMUX
D_1
D_{NL}
310A
FREQUENCY SUB-CHANNEL DATA PROCESSOR
CSI
310L
FREQUENCY SUB-CHANNEL DATA PROCESSOR
CSI
MODULATION SYMBOL STREAMS
$S_{1,1}$
$S_{1,Nc}$
$S_{1,1}$
$S_{NL,Nc}$
120C
322A
SUB-CHANNEL MIMO PROCESSOR
CSI
322N
SUB-CHANNEL MIMO PROCESSOR
CSI
324A
COMBINER
324T
COMBINER
SYMBOL VECTOR STREAMS
V_1
V_{Nt}
122A
320A
IFFT
322A
CYCLIC PREFIX GENERATOR
324A
UP-CONVERTER
124A
122T
320T
IFFT
322T
CYCLIC PREFIX GENERATOR
324T
UP-CONVERTER
124T

FIG. 3

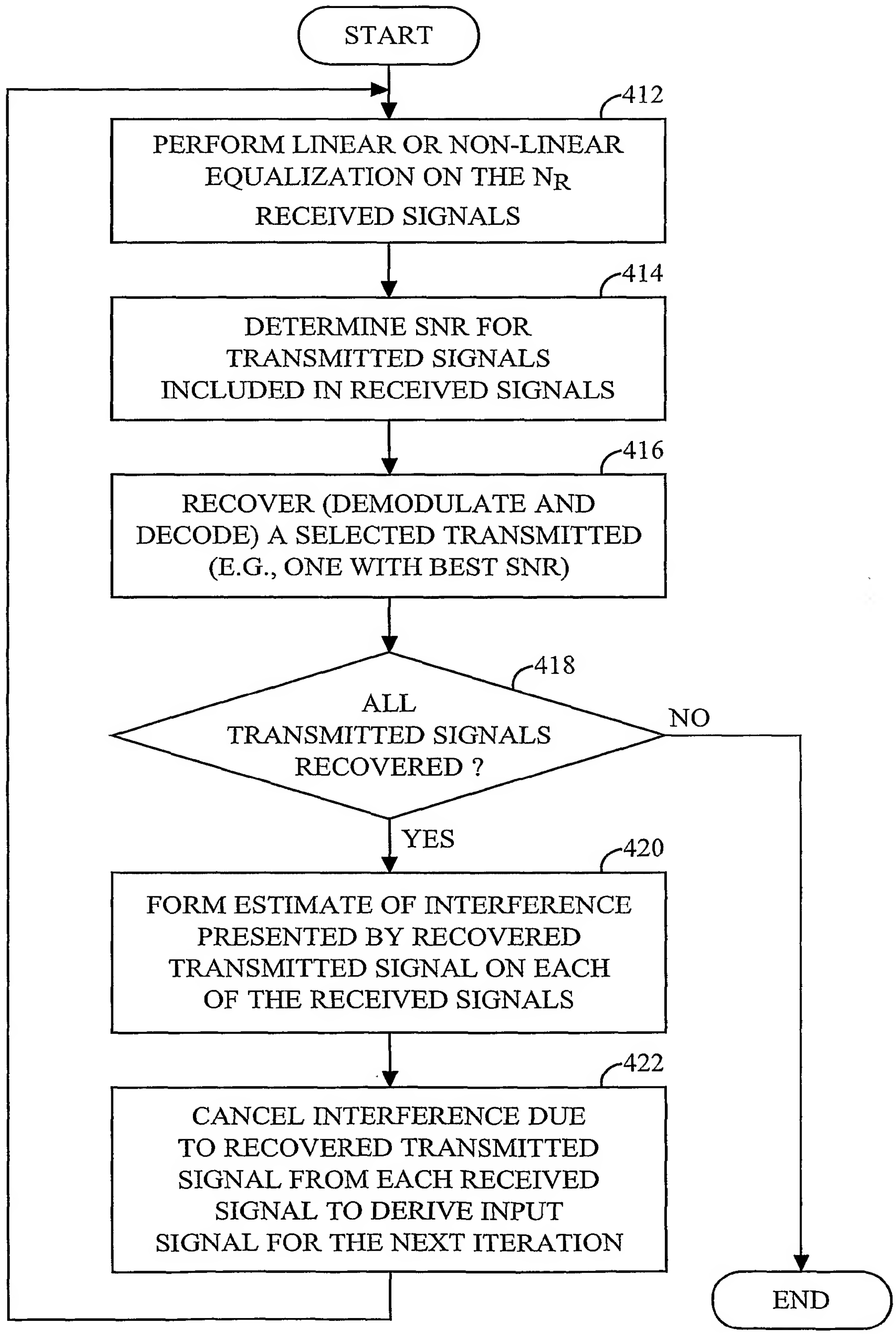
START
412
PERFORM LINEAR OR NON-LINEAR EQUALIZATION ON THE N_R RECEIVED SIGNALS
414
DETERMINE SNR FOR TRANSMITTED SIGNALS INCLUDED IN RECEIVED SIGNALS
416
RECOVER (DEMODULATE AND DECODE) A SELECTED TRANSMITTED (E.G., ONE WITH BEST SNR)
418
ALL TRANSMITTED SIGNALS RECOVERED ?
NO
YES
420
FORM ESTIMATE OF INTERFERENCE PRESENTED BY RECOVERED TRANSMITTED SIGNAL ON EACH OF THE RECEIVED SIGNALS
422
CANCEL INTERFERENCE DUE TO RECOVERED TRANSMITTED SIGNAL FROM EACH RECEIVED SIGNAL TO DERIVE INPUT SIGNAL FOR THE NEXT ITERATION
END

FIG. 4

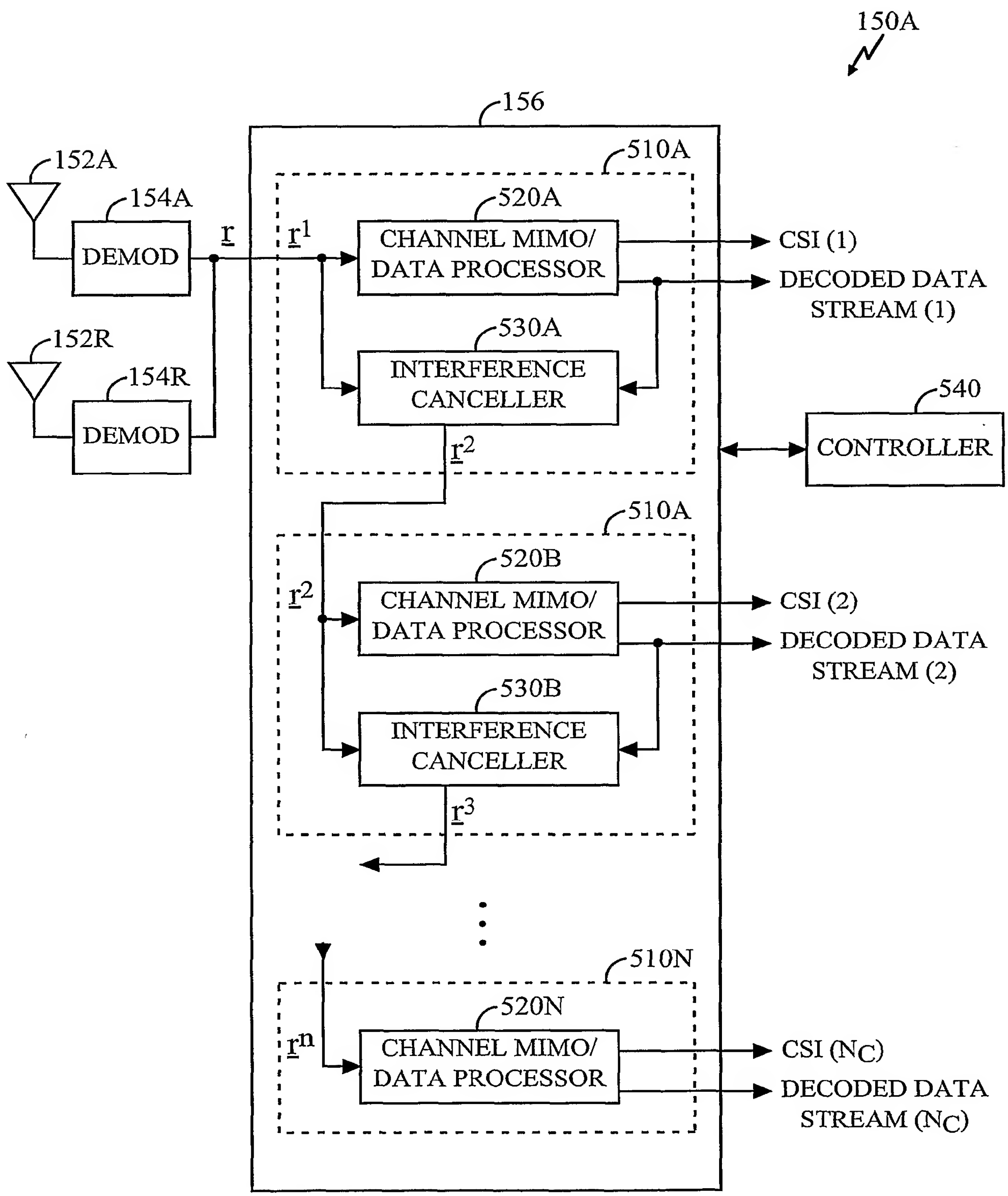
150A
156
152A
154A
DEMOD
r
152R
154R
DEMOD
510A
520A
r1
CHANNEL MIMO/ DATA PROCESSOR
530A
INTERFERENCE CANCELLER
r2
CSI (1)
DECODED DATA STREAM (1)
540
CONTROLLER
510A
520B
r2
CHANNEL MIMO/ DATA PROCESSOR
530B
INTERFERENCE CANCELLER
r3
CSI (2)
DECODED DATA STREAM (2)
510N
520N
rn
CHANNEL MIMO/ DATA PROCESSOR
CSI (NC)
DECODED DATA STREAM (NC)

FIG. 5

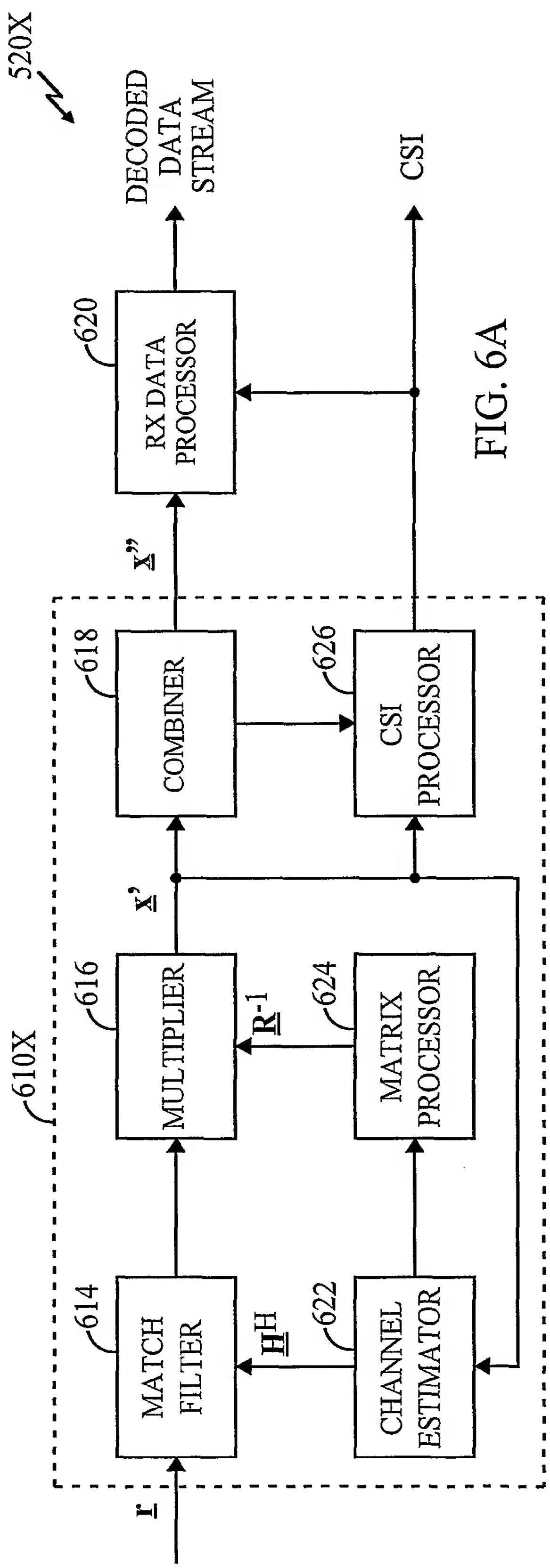
520X
610X
r
MATCH FILTER
614
$\underline{H}^H$
CHANNEL ESTIMATOR
622
MULTIPLIER
616
$\underline{R}^{-1}$
MATRIX PROCESSOR
624
x'
COMBINER
618
CSI PROCESSOR
626
x''
RX DATA PROCESSOR
620
DECODED DATA STREAM
CSI

FIG. 6A

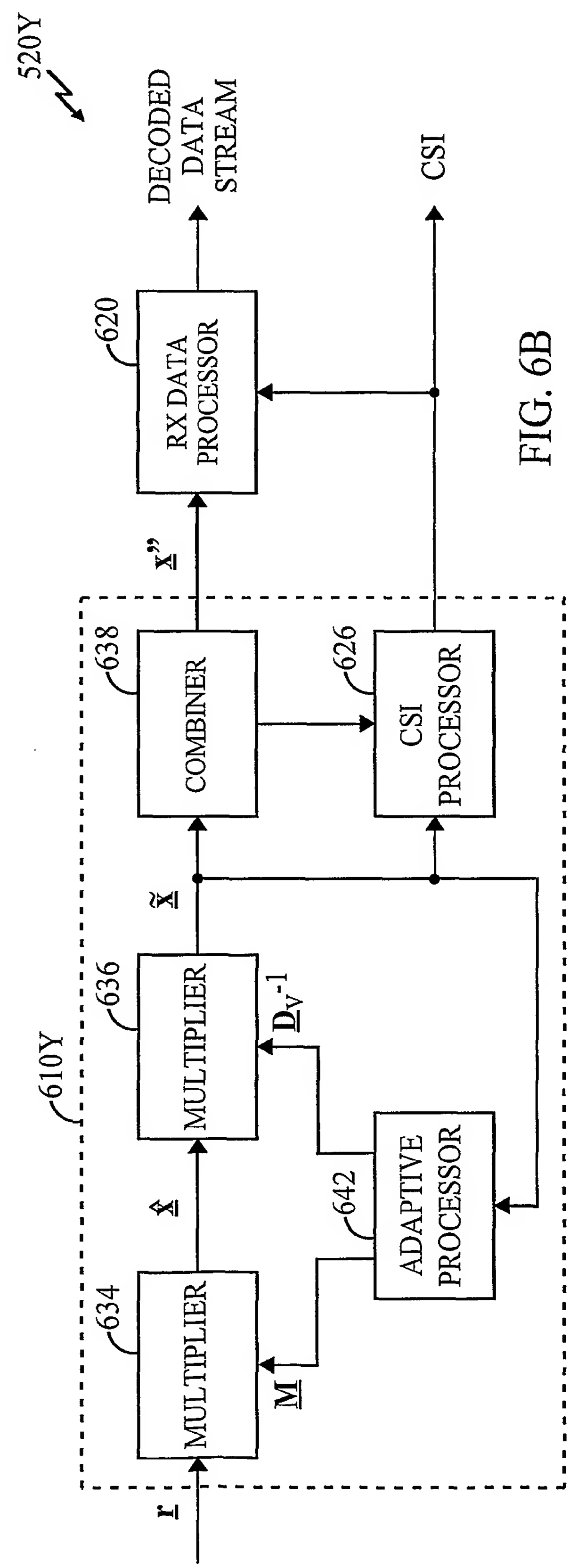
520Y
610Y
r
MULTIPLIER
634
M
ADAPTIVE PROCESSOR
642
$\hat{x}$
MULTIPLIER
636
$\underline{D}_V^{-1}$
$\tilde{x}$
COMBINER
638
CSI PROCESSOR
626
x''
RX DATA PROCESSOR
620
DECODED DATA STREAM
CSI

FIG. 6B

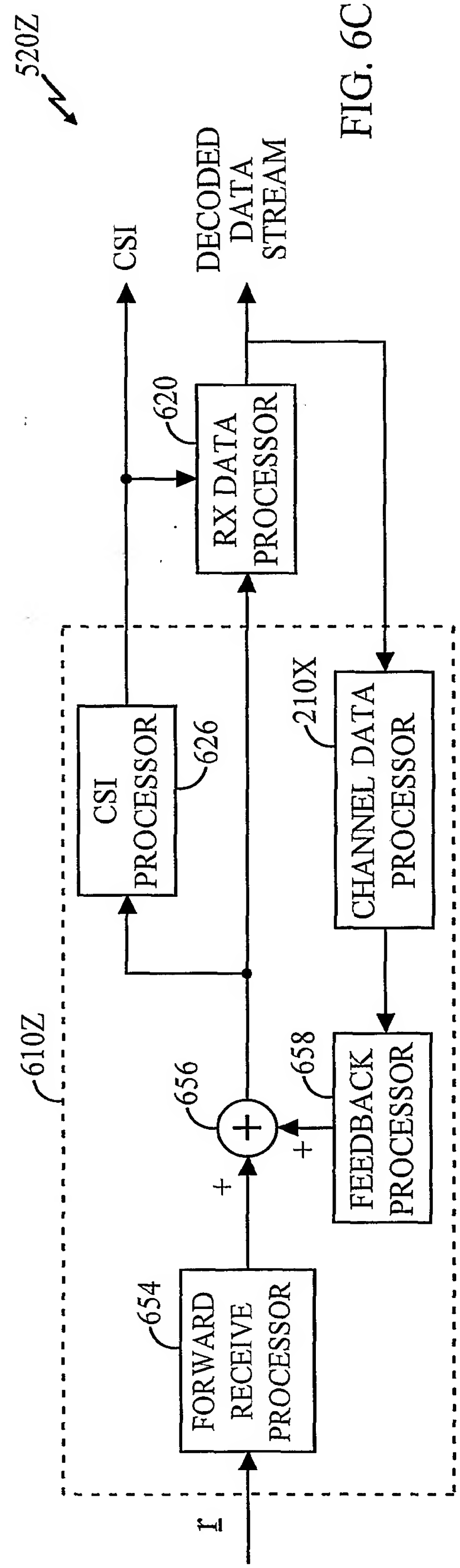

FIG. 6C

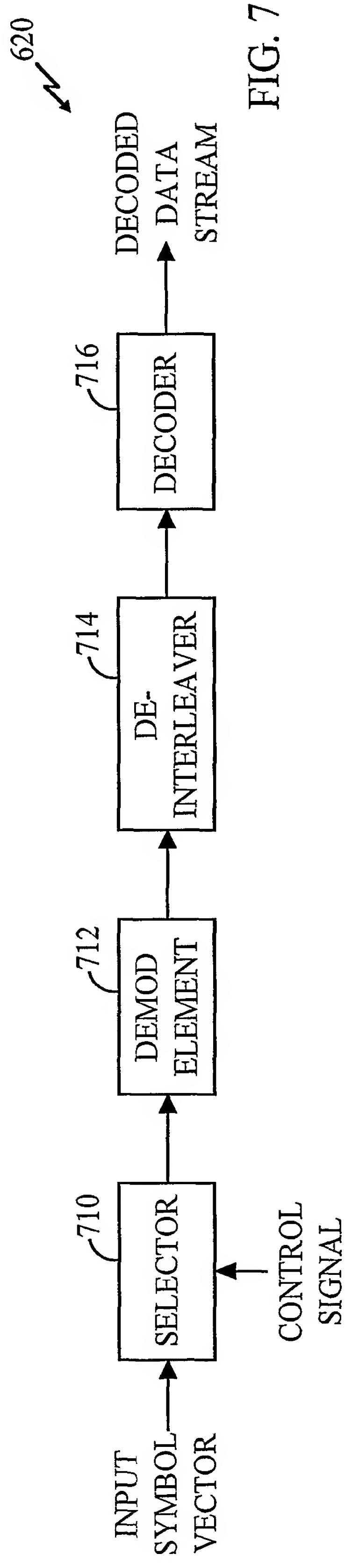

FIG. 7

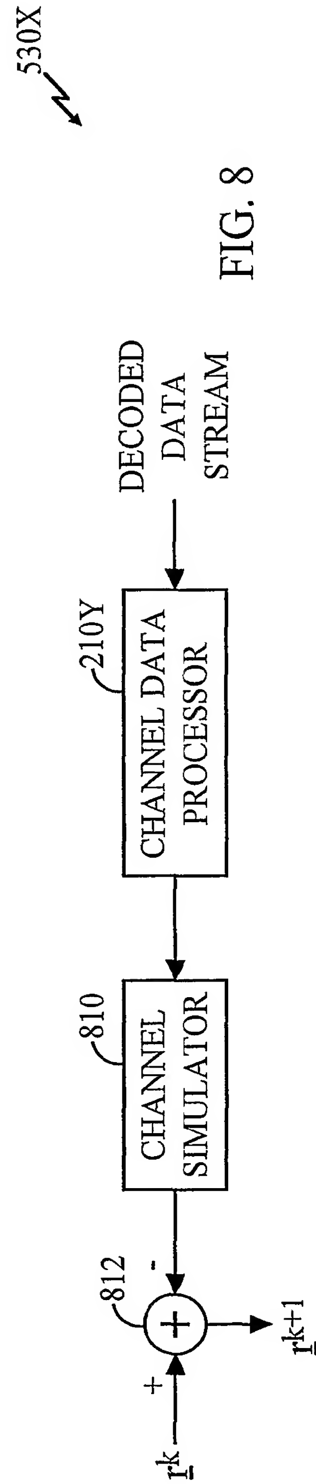

FIG. 8

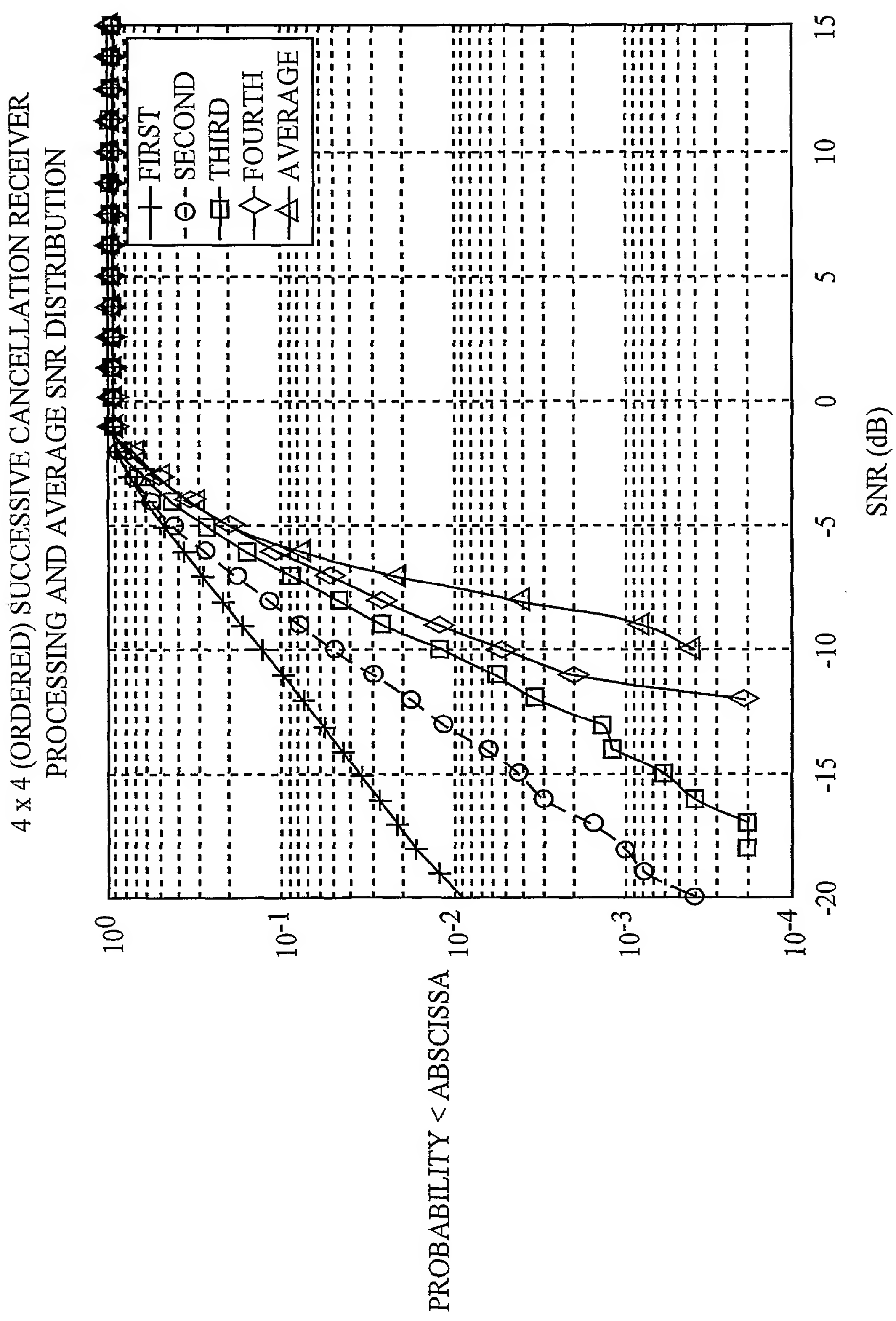
4 x 4 (ORDERED) SUCCESSIVE CANCELLATION RECEIVER
PROCESSING AND AVERAGE SNR DISTRIBUTION
FIRST
SECOND
THIRD
FOURTH
AVERAGE
PROBABILITY < ABSCISSA
10^0
10^{-1}
10^{-2}
10^{-3}
10^{-4}
-20
-15
-10
-5
0
5
10
15
SNR (dB)

FIG. 9A

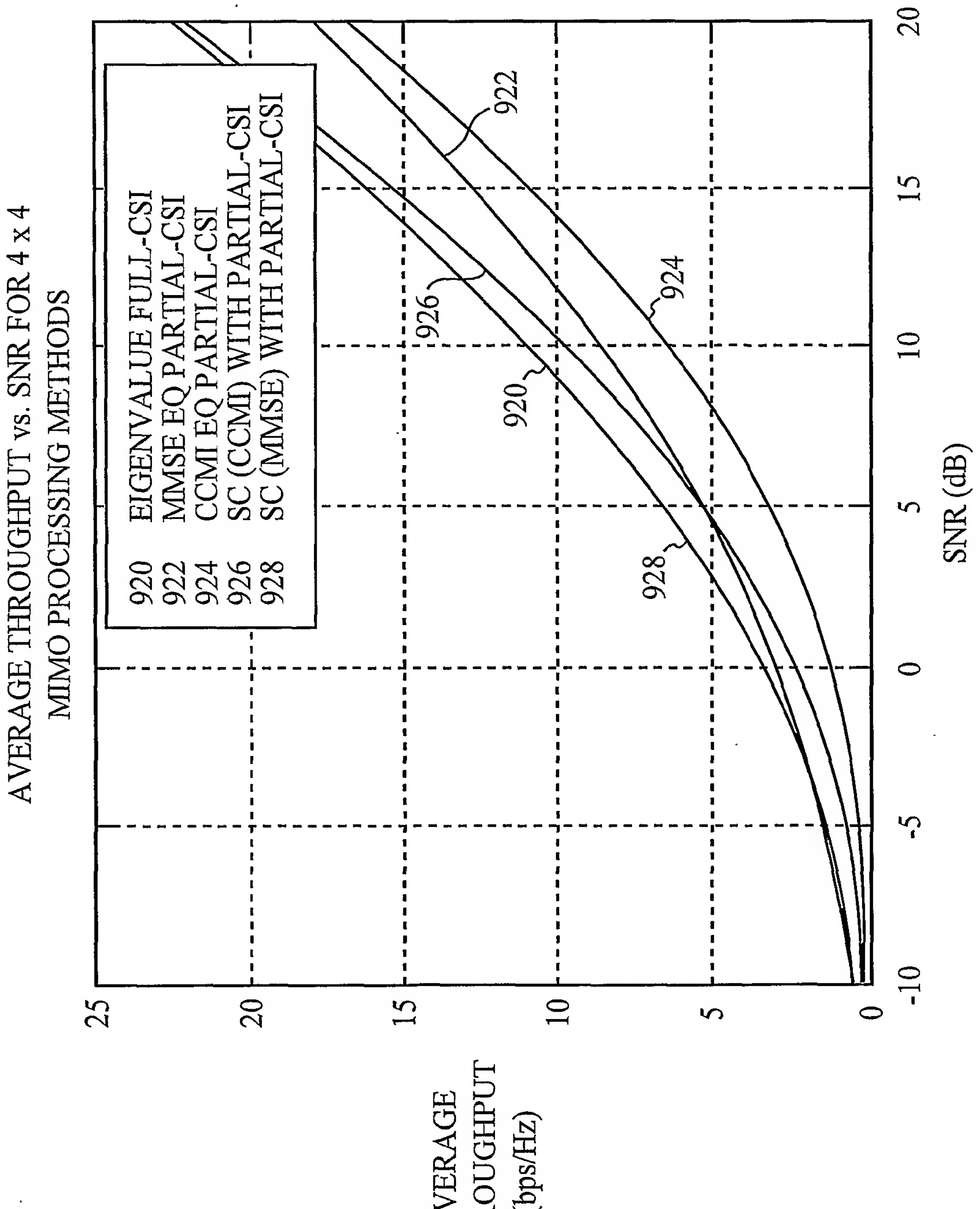
AVERAGE THROUGHPUT vs. SNR FOR 4 x 4
MIMO PROCESSING METHODS
920 EIGENVALUE FULL-CSI
922 MMSE EQ PARTIAL-CSI
924 CCMI EQ PARTIAL-CSI
926 SC (CCMI) WITH PARTIAL-CSI
928 SC (MMSE) WITH PARTIAL-CSI
AVERAGE
THROUGHPUT
(bps/Hz)
SNR (dB)
0
5
10
15
20
25
-10
-5
0
5
10
15
20
920
922
924
926
928

FIG. 9B

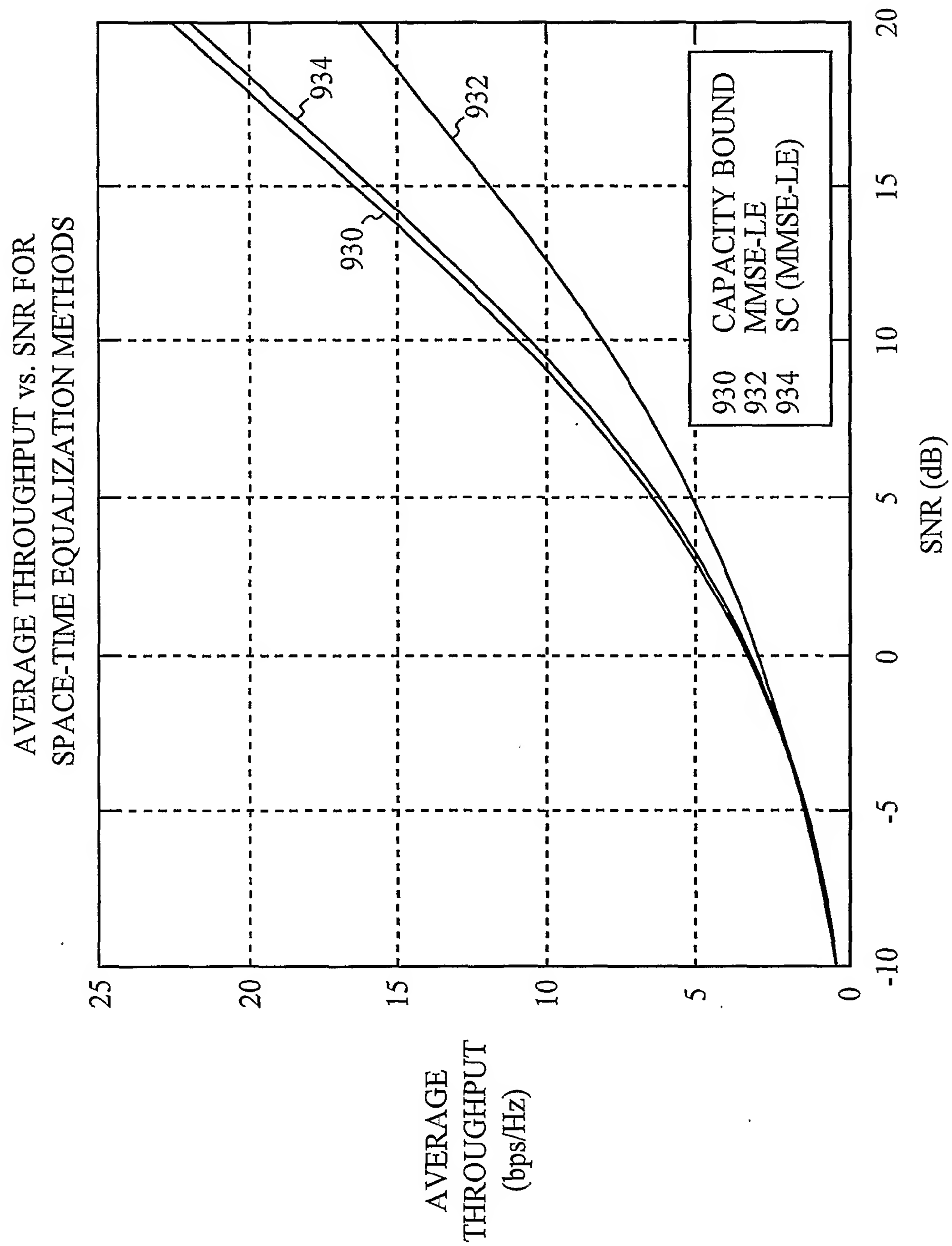
AVERAGE THROUGHPUT vs. SNR FOR
SPACE-TIME EQUALIZATION METHODS
AVERAGE
THROUGHPUT
(bps/Hz)
0
5
10
15
20
25
-10
-5
0
5
10
15
20
SNR (dB)
930
932
934
930 CAPACITY BOUND
932 MMSE-LE
934 SC (MMSE-LE)

FIG. 9C

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 02/14526

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04B7/08 H04B7/06 H04L1/06

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04B H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data, PAJ, INSPEC

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	WO 98 09381 A (UNIV LELAND STANFORD JUNIOR) 5 March 1998 (1998-03-05) abstract; claims 1-5,12; figures 6A,6B,7 page 3, line 28 -page 7, last line ---	1-7,11, 15-57
Y	EP 0 951 091 A (LUCENT TECHNOLOGIES INC) 20 October 1999 (1999-10-20) abstract; claims 1,2; figures 1-8 paragraph '0003! ---	1-7,11, 15-57

-/--

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
29 August 2002	06/09/2002

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31–70) 340–2040, Tx. 31 651 epo nl, Fax: (+31–70) 340–3016	Katruff, M

INTERNATIONAL SEARCH REPORT

Inte onal Application No
PCi/US 02/14526

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
A	JOENGREN G ET AL: "UTILIZING QUANTIZED FEEDBACK INFORMATION IN ORTHOGONAL SPACE-TIME BLOCK CODING" GLOBECOM'00. 2000 IEEE GLOBAL TELECOMMUNICATIONS CONFERENCE. SAN FRANCISCO, CA, NOV. 27 - DEC. 1, 2000, IEEE GLOBAL TELECOMMUNICATIONS CONFERENCE, NEW YORK, NY: IEEE, US, vol. 2 OF 4, 27 November 2000 (2000-11-27), pages 995-999, XP001017234 ISBN: 0-7803-6452-X abstract; figure 1 page 996, left-hand column, paragraph 1 -page 997, left-hand column, paragraph 1 ---	1,38,39, 47,48
A	US 6 131 016 A (SOLLENBERGER NELSON RAY ET AL) 10 October 2000 (2000-10-10) abstract; figure 2A column 1, line 41 -column 2, line 11; claim 1 -----	1,38,39, 47,48

INTERNATIONAL SEARCH REPORT

formation on patent family members

Inte ional Application No
PC i/US 02/14526

Patent document cited in search report		Publication date	Patent family member(s)		Publication date
WO 9809381	A	05-03-1998	AU	4238697 A	19-03-1998
			CA	2302289 A1	05-03-1998
			EP	0920738 A1	09-06-1999
			EP	0931388 A2	28-07-1999
			JP	2001505723 T	24-04-2001
			WO	9809385 A2	05-03-1998
			WO	9809381 A1	05-03-1998
			WO	9809395 A1	05-03-1998
			US	6377631 B1	23-04-2002
			US	6144711 A	07-11-2000
EP 0951091	A	20-10-1999	US	6317466 B1	13-11-2001
			EP	0951091 A2	20-10-1999
			JP	2000101667 A	07-04-2000
			US	2001050964 A1	13-12-2001
US 6131016	A	10-10-2000	CA	2245240 A1	27-02-1999

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
12 September 2003 (12.09.2003)

PCT

(10) International Publication Number
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(54) Title: DATA TRANSMISSION WITH NON-UNIFORM DISTRIBUTION OF DATA RATES FOR A MULTIPLE-INPUT MULTIPLE-OUTPUT (MIMO) SYSTEM

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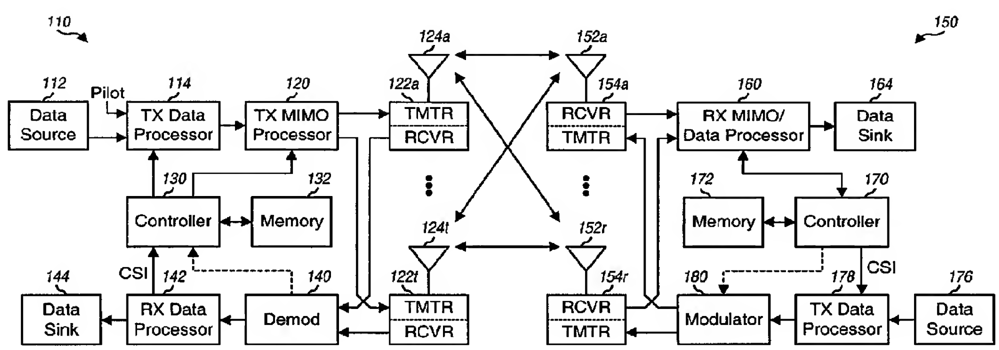

(57) Abstract: Techniques to determine data rates for a number of data streams transmitted via a number of transmission channels (or transmit antennas) in a multi-channel (e.g., MIMO) communication system. In one method, the "required" SNR for each data rate to be used is initially determined, with at least two data rates being unequal. The "effective" SNR for each data stream is also determined based on the received SNR and successive interference cancellation processing at the receiver to recover the data streams. The required SNR for each data stream is then compared against its effective SNR. The data rates are deemed to be supported if the required SNR for each data stream is less than or equal to its effective SNR. A number of sets of data rates may be evaluated, and the rate set associated with the minimum received SNR may be selected for use for the data streams.

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DATA TRANSMISSION WITH NON-UNIFORM DISTRIBUTION OF DATA RATES FOR A MULTIPLE-INPUT MULTIPLE-OUTPUT (MIMO) SYSTEM

BACKGROUND

Field

[1001] The present invention relates generally to data communication, and more specifically to techniques for determining a non-uniform distribution of data rates to be used for multiple data streams to be transmitted via multiple transmission channels of a multi-channel communication system, e.g., a multiple-input multiple-output (MIMO) system.

Background

[1002] In a wireless communication system, an RF modulated signal from a transmitter may reach a receiver via a number of propagation paths. The characteristics of the propagation paths typically vary over time due to a number of factors such as fading and multipath. To provide diversity against deleterious path effects and improve performance, multiple transmit and receive antennas may be used. If the propagation paths between the transmit and receive antennas are linearly independent (i.e., a transmission on one path is not formed as a linear combination of the transmissions on the other paths), which is generally true to at least an extent, then the likelihood of correctly receiving a data transmission increases as the number of antennas increases. Generally, diversity increases and performance improves as the number of transmit and receive antennas increases.

[1003] A multiple-input multiple-output (MIMO) communication system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels may also be referred to as a spatial subchannel (or a transmission channel) of the MIMO channel and corresponds to a dimension. The MIMO system can provide improved performance (e.g., increased transmission

capacity) if the additional dimensionalities created by the multiple transmit and receive antennas are utilized.

[1004] For a full-rank MIMO channel, where $N_S = N_T \leq N_R$, an independent data stream may be transmitted from each of the N_T transmit antennas. The transmitted data streams may experience different channel conditions (e.g., different fading and multipath effects) and may achieve different signal-to-noise-and-interference ratios (SNRs) for a given amount of transmit power. Moreover, if successive interference cancellation processing is used at the receiver to recover the transmitted data streams (described below), then different SNRs may be achieved for the data streams depending on the specific order in which the data streams are recovered. Consequently, different data rates may be supported by different data streams, depending on their achieved SNRs. Since the channel conditions typically vary with time, the data rate supported by each data stream also varies with time.

[1005] If the characteristics of the MIMO channel (e.g., the achieved SNRs for the data streams) are known at the transmitter, then the transmitter may be able to determine a particular data rate and coding and modulation scheme for each data stream such that an acceptable level of performance (e.g., one percent packet error rate) may be achieved for the data stream. However, for some MIMO systems, this information is not available at the transmitter. Instead, what may be available is very limited amount of information regarding, for example, the operating SNR for the MIMO channel, which may be defined as the expected SNR for all data streams at the receiver. In this case, the transmitter would need to determine the proper data rate and coding and modulation scheme for each data stream based on this limited information.

[1006] There is therefore a need in the art for techniques to determine a set of data rates for multiple data streams to achieve high performance when limited information is available at the transmitter for the MIMO channel.

SUMMARY

[1007] Techniques are provided herein to provide improved performance for a MIMO system when channel state information indicative of the current channel conditions is not available at the transmitter. In an aspect, a non-uniform distribution of data rates is used for the transmitted data streams. The data rates may be selected to achieve (1) a specified overall spectral efficiency with a lower minimum "received"

SNR (described below) or (2) a higher overall spectral efficiency for a specified received SNR. A specific scheme for achieving each of the above objectives is provided herein.

[1008] In a specific embodiment that may be used to achieve the first objective noted above, a method is provided for determining data rates to be used for a number of data streams to be transmitted via a number of transmission channels in a multi-channel communication system (e.g., one data stream may be transmitted over each transmit antenna in a MIMO system). In accordance with the method, the required SNR for each of a number of data rates to be used for the data streams is initially determined. At least two of the data rates are unequal. The "effective" SNR (described below) for each data stream is also determined based on the received SNR and successive interference cancellation processing at the receiver (also described below) to recover the data streams. The required SNR for each data stream is then compared against the effective SNR for the data stream. The data rates are deemed to be supported if the required SNR for each data stream is less than or equal to the effective SNR for the data stream. A number of sets of data rates may be evaluated, and the rate set associated with the minimum received SNR may be selected for use for the data streams.

[1009] In a specific embodiment that may be used to achieve the second objective noted above, a method is provided for determining data rates for a number of data streams to be transmitted via a number of transmission channels (e.g., transmit antennas) in a multi-channel (e.g., MIMO) communication system. In accordance with the method, the received SNR is initially determined. This received SNR may be specified for the system or may be estimated based on measurements at the receiver and periodically provided to the transmitter. The effective SNR for each data stream is also determined based on the received SNR and successive interference cancellation processing at the receiver. The data rate for each data stream is then determined based on the effective SNR for the data stream, with at least two of the data rates being unequal.

[1010] Various aspects and embodiments of the invention are described in further detail below. The invention further provides methods, processors, transmitter units, receiver units, base stations, terminals, systems, and other apparatuses and elements that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[1011] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[1012] FIG. 1 is a block diagram of an embodiment of a transmitter system and a receiver system in a MIMO system;

[1013] FIG. 2 is a flow diagram illustrating a successive interference cancellation receiver processing technique to process N_R received symbol streams to recover N_T transmitted symbol streams;

[1014] FIG. 3 is a flow diagram of an embodiment of a process for determining the minimum received SNR needed to support a given set of data rates;

[1015] FIG. 4 shows plots of packet error rate (PER) versus SNR for a {1, 4} MIMO system for spectral efficiencies of 1, 4/3, 5/3, and 2 bps/Hz;

[1016] FIG. 5 is a block diagram of an embodiment of a transmitter unit; and

[1017] FIG. 6 is a block diagram of an embodiment of a receiver unit capable of implementing the successive interference cancellation receiver processing technique.

DETAILED DESCRIPTION

[1018] The techniques described herein for determining a set of data rates for multiple data streams based on limited channel state information may be implemented in various multi-channel communication systems. Such multi-channel communication systems include multiple-input multiple-output (MIMO) communication systems, orthogonal frequency division multiplexing (OFDM) communication systems, MIMO systems that employ OFDM (i.e., MIMO-OFDM systems), and so on. For clarity, various aspects and embodiments are described specifically for a MIMO system.

[1019] A MIMO system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels may also be referred to as a spatial subchannel (or transmission channel) of the MIMO channel. The number of spatial subchannels is determined by the number of eigenmodes for the MIMO channel,

which in turn is dependent on a channel response matrix, $\underline{\mathbf{H}}$, that describes the response between the N_T transmit and N_R receive antennas. The elements of the channel response matrix, $\underline{\mathbf{H}}$, are composed of independent Gaussian random variables $\{h_{i,j}\}$, for $i=1, 2, \ldots N_R$ and $j=1, 2, \ldots N_T$, where $h_{i,j}$ is the coupling (i.e., the complex gain) between the j-th transmit antenna and the i-th receive antenna. For simplicity, the channel response matrix, $\underline{\mathbf{H}}$, is assumed to be full-rank (i.e., $N_S = N_T \leq N_R$), and one independent data stream may be transmitted from each of the N_T transmit antennas.

[1020] FIG. 1 is a block diagram of an embodiment of a transmitter system 110 and a receiver system 150 in a MIMO system 100.

[1021] At transmitter system 110, traffic data for a number of data streams is provided from a data source 112 to a transmit (TX) data processor 114. In an embodiment, each data stream is transmitted over a respective transmit antenna. TX data processor 114 formats, codes, and interleaves the traffic data for each data stream based on a particular coding scheme selected for that data stream to provide coded data.

[1022] The coded data for each data stream may be multiplexed with pilot data using, for example, time division multiplexing (TDM) or code division multiplexing (CDM). The pilot data is typically a known data pattern that is processed in a known manner (if at all), and may be used at the receiver system to estimate the channel response. The multiplexed pilot and coded data for each data stream is then modulated (i.e., symbol mapped) based on a particular modulation scheme (e.g., BPSK, QSPK, M-PSK, or M-QAM) selected for that data stream to provide modulation symbols. The data rate, coding, and modulation for each data stream may be determined by controls provided by a controller 130.

[1023] The modulation symbols for all data streams are then provided to a TX MIMO processor 120, which may further process the modulation symbols (e.g., for OFDM). TX MIMO processor 120 then provides N_T modulation symbol streams to N_T transmitters (TMTR) 122a through 122t. Each transmitter 122 receives and processes a respective symbol stream to provide one or more analog signals, and further conditions (e.g., amplifies, filters, and upconverts) the analog signals to provide a modulated signal suitable for transmission over the MIMO channel. N_T modulated signals from transmitters 122a through 122t are then transmitted from N_T antennas 124a through 124t, respectively.

[1024] At receiver system 150, the transmitted modulated signals are received by N_R antennas 152a through 152r, and the received signal from each antenna 152 is provided to a respective receiver (RCVR) 154. Each receiver 154 conditions (e.g., filters, amplifies, and downconverts) a respective received signal, digitizes the conditioned signal to provide samples, and further processes the samples to provide a corresponding "received" symbol stream.

[1025] An RX MIMO/data processor 160 then receives and processes the N_R received symbol streams from N_R receivers 154 based on a particular receiver processing technique to provide N_T "detected" symbol streams. The processing by RX MIMO/data processor 160 is described in further detail below. Each detected symbol stream includes symbols that are estimates of the modulation symbols transmitted for the corresponding data stream. RX MIMO/data processor 160 then demodulates, deinterleaves, and decodes each detected symbol stream to recover the traffic data for the data stream. The processing by RX MIMO/data processor 160 is complementary to that performed by TX MIMO processor 120 and TX data processor 114 at transmitter system 110.

[1026] RX MIMO processor 160 may derive an estimate of the channel response between the N_T transmit and N_R receive antennas, e.g., based on the pilot multiplexed with the traffic data. The channel response estimate may be used to perform space or space/time processing at the receiver. RX MIMO processor 160 may further estimate the signal-to-noise-and-interference ratios (SNRs) of the detected symbol streams, and possibly other channel characteristics, and provides these quantities to a controller 170. RX MIMO/data processor 160 or controller 170 may further derive an estimate of the "operating" SNR for the system, which is indicative of the conditions of the communication link. Controller 170 then provides channel state information (CSI), which may comprise various types of information regarding the communication link and/or the received data stream. For example, the CSI may comprise only the operating SNR. The CSI is then processed by a TX data processor 178, modulated by a modulator 180, conditioned by transmitters 154a through 154r, and transmitted back to transmitter system 110.

[1027] At transmitter system 110, the modulated signals from receiver system 150 are received by antennas 124, conditioned by receivers 122, demodulated by a demodulator 140, and processed by a RX data processor 142 to recover the CSI reported

by the receiver system. The reported CSI is then provided to controller 130 and used to (1) determine the data rates and coding and modulation schemes to be used for the data streams and (2) generate various controls for TX data processor 114 and TX MIMO processor 120.

[1028] Controllers 130 and 170 direct the operation at the transmitter and receiver systems, respectively. Memories 132 and 172 provide storage for program codes and data used by controllers 130 and 170, respectively.

[1029] The model for the MIMO system may be expressed as:

$$\underline{\mathbf{y}} = \underline{\mathbf{H}}\underline{\mathbf{x}} + \underline{\mathbf{n}} \ , \qquad \text{Eq (1)}$$

where $\underline{\mathbf{y}}$ is the received vector, i.e., $\underline{\mathbf{y}} = [y_1 \ y_2 \ldots \ y_{N_R}]^T$, where $\{y_i\}$ is the entry received on the i-th received antenna and $i \in \{1, \ldots, N_R\}$;

$\underline{\mathbf{x}}$ is the transmitted vector, i.e., $\underline{\mathbf{x}} = [x_1 \ x_2 \ldots \ x_{N_T}]^T$, where $\{x_j\}$ is the entry transmitted from the j-th transmit antenna and $j \in \{1, \ldots, N_T\}$;

$\underline{\mathbf{H}}$ is the channel response matrix for the MIMO channel;

$\underline{\mathbf{n}}$ is the additive white Gaussian noise (AWGN) with a mean vector of $\underline{\mathbf{0}}$ and a covariance matrix of $\underline{\mathbf{\Lambda}}_n = \sigma^2 \underline{\mathbf{I}}$, where $\underline{\mathbf{0}}$ is a vector of zeros, $\underline{\mathbf{I}}$ is the identity matrix with ones along the diagonal and zeros everywhere else, and σ^2 is the variance of the noise; and

$[.]^T$ denotes the transpose of $[.]$.

[1030] Due to scattering in the propagation environment, the N_T symbol streams transmitted from the N_T transmit antennas interfere with each other at the receiver. In particular, a given symbol stream transmitted from one transmit antenna may be received by all N_R receive antennas at different amplitudes and phases. Each received signal may then include a component of each of the N_T transmitted symbol streams. The N_R received signals would collectively include all N_T transmitted symbols streams. However, these N_T symbol streams are dispersed among the N_R received signals.

[1031] At the receiver, various processing techniques may be used to process the N_R received signals to detect the N_T transmitted symbol streams. These receiver processing techniques may be grouped into two primary categories:

- spatial and space-time receiver processing techniques (which are also referred to as equalization techniques), and
- "successive nulling/equalization and interference cancellation" receiver processing technique (which is also referred to as "successive interference cancellation" or "successive cancellation" receiver processing technique).

[1032] In general, the spatial and space-time receiver processing techniques attempt to separate out the transmitted symbol streams at the receiver. Each transmitted symbol stream may be "detected" by (1) combining the various components of the transmitted symbol stream included in the N_R received signals based on an estimate of the channel response and (2) removing (or canceling) the interference due to the other symbol streams. These receiver processing techniques attempt to either (1) decorrelate the individual transmitted symbol streams such that there is no interference from the other symbol streams or (2) maximize the SNR of each detected symbol stream in the presence of noise and interference from the other symbol streams. Each detected symbol stream is then further processed (e.g., demodulated, deinterleaved, and decoded) to recover the traffic data for the symbol stream.

[1033] The successive cancellation receiver processing technique attempts to recover the transmitted symbol streams, one at a time, using spatial or space-time receiver processing, and to cancel the interference due to each "recovered" symbol stream such that later recovered symbol streams experience less interference and may be able to achieve higher SNRs. The successive cancellation receiver processing technique may be used if the interference due to each recovered symbol stream can be accurately estimated and canceled, which requires error-free or low-error recovery of the symbol stream. The successive cancellation receiver processing technique (which is described in further detail below) generally outperforms the spatial/space-time receiver processing techniques.

[1034] For the successive cancellation receiver processing technique, the N_R received symbol streams are processed by N_T stages to successively recover one transmitted symbol stream at each stage. As each transmitted symbol stream is recovered, the interference it causes to the remaining not yet recovered symbol streams is estimated and canceled from the received symbol streams, and the "modified" symbol streams are further processed by the next stage to recover the next transmitted symbol stream. If the transmitted symbol streams can be recovered without error (or with

minimal errors) and if the channel response estimate is reasonably accurate, then cancellation of the interference due to the recovered symbol stream is effective, and the SNR of each subsequently recovered symbol stream is improved. In this way, higher performance may be achieved for all transmitted symbol streams (possibly except for the first transmitted symbol stream to be recovered).

[1035] The following terminology is used herein:

- "transmitted" symbol streams - the symbol streams transmitted from the transmit antennas;
- "received" symbol streams - the inputs to a spatial or space-time processor in the first stage of a successive interference cancellation (SIC) receiver (see FIG. 6);
- "modified" symbol streams - the inputs to the spatial or space-time processor in each subsequent stage of the SIC receiver;
- "detected" symbol streams - the outputs from the spatial processor (up to $N_T - k + 1$ symbol streams may be detected at stage *k*); and
- "recovered" symbol stream - a symbol stream that has been decoded at the receiver (only one detected symbol stream is recovered at each stage).

[1036] FIG. 2 is a flow diagram illustrating the successive cancellation receiver processing technique to process N_R received symbol streams to recover N_T transmitted symbol streams. For simplicity, the following description for FIG. 2 assumes that (1) the number of spatial subchannels is equal to the number of transmit antennas (i.e., $N_S = N_T \le N_R$) and (2) one independent data stream is transmitted from each transmit antenna.

[1037] For the first stage ($k = 1$), the receiver initially performs spatial or space-time processing on the N_R received symbol streams to attempt to separate out the N_T transmitted symbol streams (step 212). For the first stage, the spatial or space-time processing can provide N_T detected symbol streams that are estimates of the N_T (not yet recovered) transmitted symbol streams. One of the detected symbol streams is then selected (e.g., based on a particular selection scheme) and further processed. If the identity of the transmitted symbol stream to be recovered in the stage is known *a priori*, then the space or space-time processing may be performed to provide only one detected symbol stream for this transmitted symbol stream. In either case, the selected detected symbol stream is further processed (e.g., demodulated, deinterleaved, and decoded) to

obtain a decoded data stream, which is an estimate of the data stream for the transmitted symbol stream being recovered in this stage (step 214).

[1038] A determination is then made whether or not all transmitted symbol streams have been recovered (step 216). If the answer is yes, then the receiver processing terminates. Otherwise, the interference due to the just-recovered symbol stream on each of the N_R received symbol streams is estimated (step 218). The interference may be estimated by first re-encoding the decoded data stream, interleaving the re-encoded data, and symbol mapping the interleaved data (using the same coding, interleaving, and modulation schemes used at the transmitter unit for this data stream) to obtain a "remodulated" symbol stream, which is an estimate of the transmitted symbol stream just recovered. The remodulated symbol stream is then convolved by each of N_R elements in a channel response vector $\underline{\mathbf{h}}_j$ to derive N_R interference components due to the just-recovered symbol stream. The vector $\underline{\mathbf{h}}_j$ is a column of the $(N_R \times N_T)$ channel response matrix, $\underline{\mathbf{H}}$, corresponding to the j-th transmit antenna used for the just-recovered symbol stream. The vector $\underline{\mathbf{h}}_j$ includes N_R elements that define the channel response between the j-th transmit antenna and the N_R receive antennas

[1039] The N_R interference components are then subtracted from the N_R received symbol streams to derive N_R modified symbol streams (step 220). These modified symbol streams represent the streams that would have been received if the just-recovered symbol stream had not been transmitted (i.e., assuming that the interference cancellation was effectively performed).

[1040] The processing performed in steps 212 and 214 is then repeated on the N_R modified symbol streams (instead of the N_R received symbol streams) to recover another transmitted symbol stream. Steps 212 and 214 are thus repeated for each transmitted symbol stream to be recovered, and steps 218 and 220 are performed if there is another transmitted symbol stream to be recovered.

[1041] For the first stage, the input symbol streams are the N_R received symbol streams from the N_R received antennas. And for each subsequent stage, the input symbol streams are the N_R modified symbol streams from the preceding stage. The processing for each stage proceeds in similar manner. At each stage subsequent to the first stage, the symbol streams recovered in the prior stages are assumed to be cancelled,

so the dimensionality of the channel response matrix $\underline{\mathbf{H}}$ is successively reduced by one column for each subsequent stage.

[1042] The successive cancellation receiver processing thus includes a number of stages, one stage for each transmitted symbol stream to be recovered. Each stage recovers one of the transmitted symbol streams and (except for the last stage) cancels the interference due to this recovered symbol stream to derive the modified symbol streams for the next stage. Each subsequently recovered symbol stream thus experiences less interference and is able to achieve a higher SNR than without the interference cancellation. The SNRs of the recovered symbol streams are dependent on the particular order in which the symbol streams are recovered.

[1043] For the successive cancellation receiver processing, the input symbol streams for the k-th stage (assuming that the interference from the symbol streams recovered in the prior k-1 stages have been effectively canceled) may be expressed as:

$$\underline{\mathbf{y}}_k = \underline{\mathbf{H}}_k \underline{\mathbf{x}}_k + \underline{\mathbf{n}} \ , \qquad \text{Eq (2)}$$

where $\underline{\mathbf{y}}_k$ is the $N_R \times 1$ input vector for the k-th stage, i.e., $\underline{\mathbf{y}}_k = [y_1^{\ k} \ y_2^{\ k} \ldots y_{N_R}^k]^T$, where $y_i^{\ k}$ is the entry for the i-th received antenna in the k-th stage;

$\underline{\mathbf{x}}_k$ is the $(N_T - k + 1) \times 1$ transmitted vector for the k-th stage, i.e., $\underline{\mathbf{x}}_k = [x_k \ x_{k+1} \ldots x_{N_T}]^T$, where x_j is the entry transmitted from the j-th transmit antenna;

$\underline{\mathbf{H}}_k$ is the $N_R \times (N_T - k + 1)$ channel response matrix for the MIMO channel, with k-1 columns for the previously recovered symbol streams removed, i.e., $\underline{\mathbf{H}}_k = [\underline{\mathbf{h}}_k \ \underline{\mathbf{h}}_{k+1} \ \ldots \ \underline{\mathbf{h}}_{N_T}]$; and

$\underline{\mathbf{n}}$ is the additive white Gaussian noise

For simplicity, equation (2) assumes that the transmitted symbol streams are recovered in the order of the transmit antennas (i.e., the symbol stream transmitted from transmit antenna 1 is recovered first, the symbol stream transmitted from transmit antenna 2 is recovered second, and so on, and the symbol stream transmitted from transmit antenna N_T is recovered last). Equation (2) may be rewritten as:

$$\underline{\mathbf{y}}_k = \sum_{j=k}^{N_T} \underline{\mathbf{h}}_j \underline{\mathbf{x}}_j + \underline{\mathbf{n}} \quad . \qquad \text{Eq (3)}$$

[1044] The transmitted symbol stream to be recovered in stage k may be viewed as being projected at a particular angle from an interference sub-space (or plane) $\underline{\mathbf{S}}^I$. The transmitted symbol stream is dependent on (and defined by) the channel response vector $\underline{\mathbf{h}}_k$. An interference-free component of the transmitted symbol stream may be obtained by projecting the channel response vector, $\underline{\mathbf{h}}_k$, on an interference-free sub-space, which is orthogonal to the interference sub-space. This projection may be achieved by multiplying $\underline{\mathbf{h}}_k$ with a filter having a response of $\underline{\mathbf{w}}$. The filter that attains the maximum energy after the projection is the one that lies in a sub-space constructed by $\underline{\mathbf{h}}_k$ and the interference sub-space $\underline{\mathbf{S}}^I$, where $\underline{\mathbf{S}}^I = \text{span}\ (\underline{\mathbf{i}}_1\ \underline{\mathbf{i}}_2\ \ldots\ \underline{\mathbf{i}}_{N_T-k})$, $\underline{\mathbf{i}}_m^H \underline{\mathbf{i}}_n = \delta_{m,n}$, and $\{\underline{\mathbf{i}}_n\}$, for $n = 1,\ 2,\ \ldots\ N_T - k$, are orthonormal basis spanning the interference sub-space $\underline{\mathbf{S}}^I$. The average energy after the projection is given by:

$$\begin{aligned} E[|\underline{\mathbf{w}}^H \underline{\mathbf{h}}_k|^2] &= E[|\underline{\mathbf{h}}_k^H \underline{\mathbf{h}}_k|^2] - E[|\underline{\mathbf{S}}^{I^H} \underline{\mathbf{h}}_k|^2] \\ &= \frac{N_R}{N_T} - \sum_{j=1}^{N_T-k} \underline{\mathbf{i}}_j^H E[|\underline{\mathbf{h}}_k \underline{\mathbf{h}}_k^H|^2] \underline{\mathbf{i}}_j \qquad \text{Eq (4)} \\ &= \frac{N_R - N_T + k}{N_T} \quad , \end{aligned}$$

where $\underline{\mathbf{w}}^H \underline{\mathbf{h}}_k$ represents the projection of $\underline{\mathbf{h}}_k$ on the interference-free sub-space (i.e., the desired component), and

$\underline{\mathbf{S}}^{I^H} \underline{\mathbf{h}}_k$ represents the projection of $\underline{\mathbf{h}}_k$ on the interference sub-space (i.e., the interference component).

Equation (4) assumes equal transmit powers being used for the transmit antennas.

[1045] The effective SNR for the symbol stream recovered in the k-th stage, $\text{SNR}_{\text{eff}}(k)$, may be expressed as:

$$\text{SNR}_{\text{eff}}(k) = \frac{P_{tot}(N_R - N_T + k)}{\sigma^2 N_T} \quad , \qquad \text{Eq (5)}$$

where P_{tot} is the total transmit power available for data transmission, which is uniformly distributed across the N_T transmit antennas such that P_{tot} / N_T is used for each transmit antenna, and

σ^2 is the noise variance.

[1046] The received SNR for all N_R received symbol streams, SNR_{rx}, may be defined as:

$$\text{SNR}_{\text{rx}} = \frac{P_{tot} N_R}{\sigma^2} \quad . \qquad \text{Eq (6)}$$

[1047] Combining equations (5) and (6), the effective SNR for the symbol stream recovered in the k-th stage may be expressed as:

$$\text{SNR}_{\text{eff}}(k) = \left(\frac{N_R - N_T + k}{N_T N_R} \right) \text{SNR}_{\text{rx}} \quad . \qquad \text{Eq (7)}$$

The effective SNR formulation shown in equation (7) is based on several assumptions. First, it is assumed that the interference due to each recovered data stream is effectively canceled and does not contribute to the noise and interference observed by the subsequently recovered symbol streams. Second, it is assumed that no (or low) errors propagate from one stage to another. Third, an optimum filter that maximizes SNR is used to obtain each detected symbol stream. Equation (7) also provides the effective SNR in linear unit (i.e., not in log or dB unit).

[1048] As noted above, the transmitted symbol streams may experience different channel conditions and may achieve different SNRs for a given amount of transmit power. If the achieved SNR of each symbol stream is known at the transmitter, then the data rate and coding and modulation scheme for the corresponding data stream may be selected to maximize spectral efficiency while achieving a target packet error rate (PER). However, for some MIMO systems, channel state information indicative of the current channel conditions is not available at the transmitter. In this case, it is not possible to perform adaptive rate control for the data streams.

[1049] Conventionally, in some MIMO systems, data is transmitted over the N_T transmit antennas at the same data rates (i.e., uniform distribution of data rates) when channel state information is not available at the transmitter. At the receiver, the N_R received symbol streams may be processed using the successive cancellation receiver

processing technique. In one conventional scheme, the SNRs of the $(N_T - k + 1)$ detected symbol streams at each stage k are determined, and the detected symbol stream with the highest SNR is recovered in that stage. This transmission scheme with uniform distribution of data rates provides sub-optimal performance.

[1050] Techniques are provided herein to provide improved performance for a MIMO system when channel state information indicative of the current channel conditions is not available at the transmitter. In an aspect, a non-uniform distribution of data rates is used for the transmitted data streams. The data rates may be selected to achieve (1) a given or specified overall spectral efficiency with a lower minimum received SNR or (2) a higher overall spectral efficiency for a given or specified received SNR. A specific scheme for achieving each of the above objectives is provided below. It can be shown that the non-uniform distribution of data rates generally outperforms the conventional uniform distribution of data rates in many situations.

[1051] As shown in equation (7), the effective SNR of each recovered symbol stream is dependent on the particular stage at which it is recovered, as indicated by the factor "k" in the numerator in equation (7). The lowest effective SNR is achieved for the first recovered symbol stream, and the highest effective SNR is achieved for the last recovered symbol stream.

[1052] To achieve improved performance, non-uniform distribution of data rates may be used for the data streams transmitted on different antennas (i.e., different spectral efficiencies may be assigned to different transmit antennas), depending on their effective SNRs. At the receiver, the transmitted data streams may be recovered in an ascending order of data rates. That is, the data stream with the lowest data rate is recovered first, the data stream with the next higher data rate is recovered second, and so on, and the data stream with the highest data rate is recovered last.

[1053] The data rates to be used for the data streams may be determined by taking into account various considerations. First, earlier recovered symbol streams achieve lower effective SNRs, as shown in equation (7), and further suffer from lower diversity order. In fact, the diversity order at stage k may be given as $(N_R - N_T + k)$. Moreover, decoding errors from earlier recovered symbol streams propagate to later recovered symbol streams and can affect the effective SNRs of these subsequently recovered symbol streams. The data rates for earlier recovered symbol streams may thus be selected to achieve high confidence in the recovery of these symbol streams and to

reduce or limit the error propagation (EP) effect on later recovered symbol streams. Second, the later recovered symbol streams may be more vulnerable to errors if they are designated to support larger spectral efficiencies, even though they may be able to achieve higher effective SNRs.

[1054] Various schemes may be implemented to (1) determine the minimum received SNR needed to support a given distribution of data rates (or spectral efficiencies) or, (2) determine the distribution of spectral efficiencies that attains the best performance for a given received SNR. One specific scheme for each of these objectives is described below.

[1055] FIG. 3 is a flow diagram of an embodiment of a process 300 for determining the minimum received SNR needed to support a given set of data rates. This set of data rates is denoted as $\{r_k\}$, for $k=1, 2, \ldots N_T$, and are ordered such that $r_1 \le r_2 \ldots \le r_{N_T}$. The data rates in set $\{r_k\}$ are to be used for the N_T data streams to be transmitted from the N_T transmit antennas.

[1056] Initially, the SNR required at the receiver to support each data rate (or spectral efficiency) in set $\{r_k\}$ is determined (step 312). This may be achieved by using a look-up table of required SNR versus spectral efficiency. The required SNR for a given spectral efficiency may be determined (e.g., using computer simulation) based on an assumption that a single data stream is transmitted over a $\{1, N_R\}$ single-input multiple-output (SIMO) channel, and is further determined for a particular target PER (e.g., 1% PER). The required SNR for a data stream with data rate r_k is denoted as $\mathrm{SNR}_{\mathrm{req}}(r_k)$. A set of N_T required SNRs is obtained in step 312 for the N_T data streams.

[1057] The N_T data rates in set $\{r_k\}$ are associated with N_T SNRs required at the receiver to achieve the target PER (e.g., as determined from the look-up table). These N_T data rates are also associated with N_T effective SNRs that may be achieved at the receiver based on a particular received SNR using successive interference cancellation processing at the receiver, as shown in equation (7). The data rates in set $\{r_k\}$ are deemed to be supported if the N_T required SNRs are at or below the corresponding effective SNRs. Visually, the N_T required SNRs may be plotted versus data rates and connected together by a first line, and the N_T effective SNRs may also be plotted versus data rates and connected together by a second line. The data rates in set $\{r_k\}$ are then deemed to be supported if no part of the first line is above the second line.

[1058] The margin for a given data rate may be defined as the difference between the effective SNR and the required SNR for the data rate, i.e., $\text{margin}(k) = \text{SNR}_{\text{eff}}(r_k) - \text{SNR}_{\text{req}}(r_k)$. The data rates in set $\{r_k\}$ may also be deemed to be supported if the margin for each data rate is zero or greater.

[1059] The effective SNRs for the data streams are dependent on the received SNR, and may be derived from the received SNR as shown in equation (7). The minimum received SNR needed to support the N_T data rates in set $\{r_k\}$ is the received SNR that results in the effective SNR of at least one data rate being equal to the required SNR (i.e., zero margin). Depending on the specific data rates included in set $\{r_k\}$, the minimum margin (of zero) may be achieved for any one of the N_T data rates in the set.

[1060] For the first iteration, the minimum margin is assumed to be achieved by the last recovered data stream, and the index variable λ is set to N_T (i.e., $\lambda = N_T$) (step 314). The effective SNR for the λ-th recovered data stream is then set equal to its required SNR (i.e., $\text{SNR}_{\text{eff}}(\lambda) = \text{SNR}_{\text{req}}(\lambda)$) (step 316). The received SNR is next determined based on the effective SNR of $\text{SNR}_{\text{eff}}(\lambda)$ for the λ-th recovered data stream, using equation (7) (step 318). For the first iteration when $\lambda = N_T$, the received SNR may be determined using equation (7) with $k = N_T$, which may then be expressed as:

$$\text{SNR}_{\text{rx}} = N_T \cdot \text{SNR}_{\text{eff}}(N_T) \quad . \qquad \text{Eq (8)}$$

The effective SNR of each remaining data stream is then determined based on the received SNR computed in step 318 and using equation (7), for $k = 1, 2, \ldots N_T - 1$ (step 320). A set of N_T effective SNRs is obtained by step 320 for the N_T data streams.

[1061] The required SNR for each data rate in set $\{r_k\}$ is then compared against the effective SNR for the data rate (step 322). A determination is next made whether or not the data rates in set $\{r_k\}$ are supported by the received SNR determined in step 318 (step 324). In particular, if the required SNR for each of the N_T data rates is less than or equal to the effective SNR for that data rate, then the data rates in set $\{r_k\}$ are deemed to be supported by the received SNR and success is declared (step 326). Otherwise, if any one of the N_T data rates exceeds the effective SNR for the data rate, then the data

rates in set $\{r_k\}$ are deemed to not be supported by the received SNR. In this case, the variable λ is decremented (i.e., $\lambda = \lambda - 1$, so that $\lambda = N_T - 1$ for the second iteration) (step 328). The process then returns to step 316 to determine the set of effective SNRs for the data rates in set $\{r_k\}$ under the assumption that the minimum margin is achieved for the second to last recovered data stream. As many iterations as necessary may be performed until success is declared in step 326. The received SNR determined in step 318 for the iteration that results in the declaration of success is then the minimum received SNR needed to support the data rates in set $\{r_k\}$.

[1062] The process shown in FIG. 3 may also be used to determine whether or not a given set of data rates is supported by a given received SNR. This received SNR may correspond to the operating SNR, SNR_{op}, which may be the average or expected (but not necessarily the instantaneous) received SNR at the receiver. The operating SNR may be determined based on measurements at the receiver and may be periodically provided to the transmitter. Alternatively, the operating SNR may be an estimate of the MIMO channel in which the transmitter is expected to operate. In any case, the received SNR is given or specified for the MIMO system.

[1063] Referring to FIG. 3, to determine whether or not the given set of data rates is supported by the given received SNR, the required SNR for each data rate may be determined initially (step 312). A set of N_T required SNRs is obtained in step 312 for the N_T data streams. Steps 314, 316, and 318 may be skipped, since the received SNR is already given. The effective SNR of each data stream is then determined based on the given received SNR and using equation (7), for $k = 1, 2, \ldots N_T$ (step 320). A set of N_T effective SNRs is obtained in step 320 for the N_T data streams.

[1064] The required SNR for each data rate in set $\{r_k\}$ is then compared against the effective SNR for that data rate (step 322). A determination is next made whether or not the data rates in set $\{r_k\}$ are supported by the received SNR. If the required SNR for each of the N_T data rates is less than or equal to the effective SNR for that data rate, then the data rates in set $\{r_k\}$ are deemed to be supported by the received SNR, and success is declared (step 326). Otherwise, if the required SNR for any one of the N_T data rates exceeds the effective SNR for the data rate, then the data rates in set $\{r_k\}$ are deemed to not be supported by the received SNR, and failure is declared.

[1065] For clarity, an example is described below for a {2, 4} MIMO system with two transmit antennas (i.e., $N_T = 2$) and four received antennas (i.e., $N_R = 4$) and designated to support an overall spectral efficiency of 3 bits per second per Hertz (bps/Hz). For this example, two sets of data rates are evaluated. The first set includes data rates corresponding to 1 bps/Hz and 2 bps/Hz, and the second set includes data rates corresponding to 4/3 bps/Hz and 5/3 bps/Hz. The performance of each rate set is determined (e.g., based on the process shown in FIG. 3) and compared against one another.

[1066] FIG. 4 shows plots of PER versus SNR for a {1, 4} MIMO system for spectral efficiencies of 1 bps/Hz, 4/3 bps/Hz, 5/3 bps/Hz, and 2 bps/Hz. These plots may be generated by computer simulation or some other means, as is known in the art. A MIMO system is typically designated to operate at a particular target PER. In this case, the SNR required to achieve the target PER for each spectral efficiency may be determined and stored in a look-up table. For example, if the target PER is 1%, then values of -2.0 dB, 0.4 dB, 3.1 dB, and 3.2 dB may be stored in the look-up table for spectral efficiencies of 1, 4/3, 5/3, and 2 bps/Hz, respectively.

[1067] For the first rate set, the required SNRs for data streams 1 and 2 with spectral efficiencies of 1 and 2 bps/Hz, respectively, may be determined (step 312 in FIG. 3) using plots 412 and 418 in FIG. 4, as follows:

$$\mathrm{SNR}_{\mathrm{req}}(1) = -2.0\ \mathrm{dB}\text{, for data stream 1 with spectral efficiency of 1 bps/Hz, and}$$

$$\mathrm{SNR}_{\mathrm{req}}(2) = 3.2\ \mathrm{dB}\text{, for data stream 2 with spectral efficiency of 2 bps/Hz.}$$

The effective SNR of data stream 2 (which is recovered last and under the assumption that the interference from data stream 1 was effectively cancelled) is then set to its required SNR (step 316), as follows:

$$\mathrm{SNR}_{\mathrm{eff}}(2) = \mathrm{SNR}_{\mathrm{req}}(2) = 3.2\ \mathrm{dB}\ .$$

The received SNR is then determined based on equation (8) (step 318), as follows:

$$\mathrm{SNR}_{\mathrm{rx}} = 2 \cdot \mathrm{SNR}_{\mathrm{req}}(2)\ , \qquad \text{for linear unit, or}$$

$$\mathrm{SNR}_{\mathrm{rx}} = \mathrm{SNR}_{\mathrm{req}}(2) + 3.0\ \mathrm{dB} = 6.2\ \mathrm{dB}\ , \qquad \text{for log unit.}$$

[1068] The effective SNR of each remaining data stream (i.e., data stream 1) is next determined based on equation (7) (step 320), as follows:

$$\mathrm{SNR}_{\mathrm{eff}}(1) = 3/8 \cdot \mathrm{SNR}_{\mathrm{rx}} \ , \qquad \text{for linear unit, or}$$

$$\mathrm{SNR}_{\mathrm{eff}}(1) = \mathrm{SNR}_{\mathrm{rx}} - 4.3\,\mathrm{dB} = 1.9\,\mathrm{dB} \ , \qquad \text{for log unit.}$$

[1069] The effective and required SNRs for each data rate in the first rate set are given in columns 2 and 3 in Table 1. The margin for each data rate is also determined and given in the last row in Table 1.

Table 1

	First rate set		Second rate set		Unit
Data stream	1	2	1	2	
Spectral efficiency	1	2	4/3	5/3	bps/Hz
$\mathrm{SNR}_{\mathrm{eff}}$	1.9	3.2	1.8	3.1	dB
$\mathrm{SNR}_{\mathrm{req}}$	-2.0	3.2	0.4	3.1	dB
margin	3.9	0.0	1.4	0.0	dB

[1070] The required SNRs for data stream 1 and 2 are then compared against the effective SNRs for these data streams (step 322). Since $\mathrm{SNR}_{\mathrm{req}}(2) = \mathrm{SNR}_{\mathrm{eff}}(2)$ and $\mathrm{SNR}_{\mathrm{req}}(1) < \mathrm{SNR}_{\mathrm{eff}}(1)$, this set of data rates is supported by a minimum received SNR of 6.2 dB.

[1071] Since the first rate set is deemed to be supported by the first iteration through the process shown in FIG. 3, no additional iterations need to be performed. However, had this first rate set not been supported by a received SNR of 6.2 dB (e.g., if the required SNR for data stream 1 turned out to be greater than 1.9 dB), then another iteration would be performed whereby the received SNR is determined based on $\mathrm{SNR}_{\mathrm{req}}(1)$ and would be greater than 6.2 dB.

[1072] For the second rate set, the required SNRs for data streams 1 and 2 with spectral efficiencies of 4/3 and 5/3 bps/Hz, respectively, may be determined using plots 414 and 416 in FIG. 4, as follows:

$\mathrm{SNR}_{\mathrm{req}}(1) = 0.4\,\mathrm{dB}$, for data stream 1 with spectral efficiency of 4/3 bps/Hz, and

$$\mathrm{SNR}_{req}(2) = 3.1\,\mathrm{dB}\,, \quad \text{for data stream 2 with spectral efficiency of 5/3 bps/Hz.}$$

The effective SNR of data stream 2 is then set to its required SNR. The received SNR is then determined based on equation (8), as follows:

$$\mathrm{SNR}_{rx} = \mathrm{SNR}_{req}(2) + 3.0\,\mathrm{dB} = 6.1\,\mathrm{dB}\,, \qquad \text{for log unit.}$$

[1073] The effective SNR of each remaining data rate (i.e., data rate 1) is next determined based on equation (7), as follows:

$$\mathrm{SNR}_{eff}(1) = \mathrm{SNR}_{rx} - 4.3\,\mathrm{dB} = 1.8\,\mathrm{dB}\,, \qquad \text{for log unit.}$$

[1074] The effective and required SNRs for each data rate in the second rate set are given in columns 4 and 5 in Table 1.

[1075] The effective SNRs of data streams 1 and 2 are then compared against their required SNRs. Again, since $\mathrm{SNR}_{req}(2) = \mathrm{SNR}_{eff}(2)$ and $\mathrm{SNR}_{req}(1) < \mathrm{SNR}_{eff}(1)$, this set of data rates is supported by a minimum received SNR of 6.1 dB.

[1076] The above description is for a "vertical" successive interference cancellation scheme whereby one data stream is transmitted from each transmit antenna and, at the receiver, one data stream is recovered at each stage of the successive interference cancellation receiver by processing the stream from one transmit antenna. The plots in FIG. 4 and the look-up table are derived for this vertical scheme.

[1077] The techniques described herein may also be used for a "diagonal" successive interference cancellation scheme whereby each data stream is transmitted from multiple (e.g., all N_T) transmit antennas (and possibly across multiple frequency bins). At the receiver, the symbols from one transmit antenna may be detected at each stage of the successive interference cancellation receiver, and each data stream may then be recovered from the symbols detected from multiple stages. For the diagonal scheme, another set of plots and another look-up table may be derived and used. The techniques described herein may also be used for other ordering schemes, and this is within the scope of the invention.

[1078] For the above example, it can be shown that, for the diagonal successive interference cancellation scheme, the minimum received SNR needed to support a uniform distribution of data rates (i.e., spectral efficiency of 1.5 bps/Hz on each of the

two data streams) is approximately 0.6 dB higher than that needed for the second rate set (i.e., spectral efficiencies of 4/3 and 5/3). This gain is achieved without severely complicating the system design.

[1079] In order to reduce the minimum received SNR needed to achieve the target PER for a given overall spectral efficiency, the last recovered data stream may be assigned with the smallest possible spectral efficiency that does not violate the no error propagation condition for any of the prior recovered data streams. If the spectral efficiency of the last recovered data stream is reduced, then the spectral efficiency of one or more prior recovered data streams needs to be increased accordingly to achieve the given overall spectral efficiency. The increased spectral efficiency for the earlier recovered data streams would then result in higher required SNRs. If the spectral efficiency of any one of the earlier recovered data streams is increased too high, then the minimum received SNR is determined by the required SNR for this data stream and not by the last recovered data stream (which is the case for the uniform distribution of data rates).

[1080] In the above example, the second rate set needs a smaller received SNR because the later recovered data stream 2 is assigned a smaller spectral efficiency that does not violate the no error propagation condition for the first recovered data stream 1. For the first rate set, the spectral efficiency assigned to data stream 1 is too conservative so that, while it assures no error propagation, it also hurts the overall performance by forcing a higher spectral efficiency to be assigned to data stream 2. In comparison, the second rate set assigns a more realistic spectral efficiency to data stream 1 that still assures no error propagation (albeit with less confidence in comparison to the first rate set). As shown in Table 1, the margin for data stream 1 for the first rate set is 3.9 dB while the margin for data stream 1 for the second rate set is 1.4 dB.

[1081] The techniques described herein may also be used to determine a set of data rates that maximizes the overall spectral efficiency for a given received SNR (which may be the operating SNR for the MIMO system). In this case, a set of effective SNRs may be initially determined for the N_T data streams based on the given received SNR and using equation (7). For each effective SNR in the set, the highest spectral efficiency that may be supported by this effective SNR for the target PER is then determined. This may be achieved by using another look-up table that stores values for spectral efficiency versus effective SNR. A set of N_T spectral efficiencies is obtained

for the set of N_T effective SNRs. A set of data rates corresponding to this set of N_T spectral efficiencies is then determined and may be used for the N_T data streams. This rate set maximizes the overall spectral efficiency for the given received SNR.

[1082] In the description above, the effective SNRs of the data streams are determined based on the received SNR and using equation (7). This equation includes various assumptions, as noted above, which are generally true (to a large extent) for typically MIMO systems. Moreover, equation (7) is also derived based on the use of successive interference cancellation processing at the receiver. A different equation or a look-up table may also be used to determine the effective SNRs of the data streams for different operating conditions and/or different receiver processing techniques, and this is within the scope of the invention.

[1083] For simplicity, the data rate determination has been described specifically for a MIMO system. These techniques may also be used for other multi-channel communication systems.

[1084] A wideband MIMO system may experience frequency selective fading, which is characterized by different amounts of attenuation across the system bandwidth. This frequency selective fading causes inter-symbol interference (ISI), which is a phenomenon whereby each symbol in a received signal acts as distortion to subsequent symbols in the received signal. This distortion degrades performance by impacting the ability to correctly detect the received symbols.

[1085] OFDM may be used to combat ISI and/or for some other considerations. An OFDM system effectively partitions the overall system bandwidth into a number of (N_F) frequency subchannels, which may also be referred to as subbands or frequency bins. Each frequency subchannel is associated with a respective subcarrier on which data may be modulated. The frequency subchannels of the OFDM system may also experience frequency selective fading, depending on the characteristics (e.g., the multipath profile) of the propagation path between the transmit and receive antennas. Using OFDM, the ISI due to frequency selective fading may be combated by repeating a portion of each OFDM symbol (i.e., appending a cyclic prefix to each OFDM symbol), as is known in the art.

[1086] For a MIMO system that utilizes OFDM (i.e., a MIMO-OFDM system), N_F frequency subchannels are available on each of the N_S spatial subchannels for data transmission. Each frequency subchannel of each spatial subchannel may be referred to

as a transmission channel, and $N_F \cdot N_S$ transmission channels are available for data transmission between the N_T transmit antennas and N_R receive antennas. The data rate determination described above may be performed for the set of N_T transmit antennas, similar to that described above for the MIMO system. Alternatively, the data rate determination may be performed independently for the set of N_T transmit antennas for each of the N_F frequency subchannels

Transmitter System

[1087] FIG. 5 is a block diagram of a transmitter unit 500, which is an embodiment of the transmitter portion of transmitter system 110 in FIG. 1. In this embodiment, a separate data rate and coding and modulation scheme may be used for each of the N_T data streams to be transmitted on the N_T transmit antennas (i.e., separate coding and modulation on a per-antenna basis). The specific data rate and coding and modulation schemes to be used for each transmit antenna may be determined based on controls provided by controller 130, and the data rates may be determined as described above.

[1088] Transmitter unit 500 includes (1) a TX data processor 114a that receives, codes, and modulates each data stream in accordance with a separate coding and modulation scheme to provide modulation symbols and (2) a TX MIMO processor 120a that may further process the modulation symbols to provide transmission symbols if OFDM is employed. TX data processor 114a and TX MIMO processor 120a are one embodiment of TX data processor 114 and TX MIMO processor 120, respectively, in FIG. 1.

[1089] In the specific embodiment shown in FIG. 5, TX data processor 114a includes a demultiplexer 510, N_T encoders 512a through 512t, N_T channel interleavers 514a through 514t, and N_T symbol mapping elements 516a through 516t, (i.e., one set of encoder, channel interleaver, and symbol mapping element for each transmit antenna). Demultiplexer 510 demultiplexes the traffic data (i.e., the information bits) into N_T data streams for the N_T transmit antennas to be used for data transmission. The N_T data streams may be associated with different data rates, as determined by the rate control. Each data stream is provided to a respective encoder 512.

[1090] Each encoder 512 receives and codes a respective data stream based on the specific coding scheme selected for that data stream to provide coded bits. The coding increases the reliability of the data transmission. The coding scheme may include any

combination of cyclic redundancy check (CRC) coding, convolutional coding, Turbo coding, block coding, and so on. The coded bits from each encoder 512 are then provided to a respective channel interleaver 514, which interleaves the coded bits based on a particular interleaving scheme. The interleaving provides time diversity for the coded bits, permits the data to be transmitted based on an average SNR for the transmission channels used for the data stream, combats fading, and further removes correlation between coded bits used to form each modulation symbol.

[1091] The coded and interleaved bits from each channel interleaver 514 are provided to a respective symbol mapping element 516, which maps these bits to form modulation symbols. The particular modulation scheme to be implemented by each symbol mapping element 516 is determined by the modulation control provided by controller 130. Each symbol mapping element 516 groups sets of q_j coded and interleaved bits to form non-binary symbols, and further maps each non-binary symbol to a specific point in a signal constellation corresponding to the selected modulation scheme (e.g., QPSK, M-PSK, M-QAM, or some other modulation scheme). Each mapped signal point corresponds to an M_j-ary modulation symbol, where M_j corresponds to the specific modulation scheme selected for the j-th transmit antenna and $M_j = 2^{q_j}$. Symbol mapping elements 516a through 516t then provide N_T streams of modulation symbols.

[1092] In the specific embodiment shown in FIG. 5, TX MIMO processor 120a includes N_T OFDM modulators, with each OFDM modulator including an inverse Fourier transform (IFFT) unit 522 and a cyclic prefix generator 524. Each IFFT 522 receives a respective modulation symbol stream from a corresponding symbol mapping element 516. Each IFFT 522 groups sets of N_F modulation symbols to form corresponding modulation symbol vectors, and converts each modulation symbol vector into its time-domain representation (which is referred to as an OFDM symbol) using the inverse fast Fourier transform. IFFT 522 may be designed to perform the inverse transform on any number of frequency subchannels (e.g., 8, 16, 32, … , N_F, …). For each OFDM symbol, cyclic prefix generator 524 repeats a portion of the OFDM symbol to form a corresponding transmission symbol. The cyclic prefix ensures that the transmission symbol retains its orthogonal properties in the presence of multipath delay spread, thereby improving performance against deleterious path effects such as channel dispersion caused by frequency selective fading. Cyclic prefix generator 524 then

provides a stream of transmission symbols to an associated transmitter 122. If OFDM is not employed, then TX MIMO processor 120a simply provides the modulation symbol stream from each symbol mapping element 516 to the associated transmitter 122.

[1093] Each transmitter 122 receives and processes a respective modulation symbol stream (for MIMO without OFDM) or transmission symbol stream (for MIMO with OFDM) to generate a modulated signal, which is then transmitted from the associated antenna 124.

[1094] Other designs for the transmitter unit may also be implemented and are within the scope of the invention.

[1095] The coding and modulation for MIMO systems with and without OFDM are described in further detail in the following U.S. patent applications:

- U.S. Patent Application Serial No. 09/993,087, entitled "Multiple-Access Multiple-Input Multiple-Output (MIMO) Communication System," filed November 6, 2001;
- U.S. Patent Application Serial No. 09/854,235, entitled "Method and Apparatus for Processing Data in a Multiple-Input Multiple-Output (MIMO) Communication System Utilizing Channel State Information," filed May 11, 2001;
- U.S. Patent Application Serial Nos. 09/826,481 and 09/956,449, both entitled "Method and Apparatus for Utilizing Channel State Information in a Wireless Communication System," respectively filed March 23, 2001 and September 18, 2001;
- U.S. Patent Application Serial No. 09/776,075, entitled "Coding Scheme for a Wireless Communication System," filed February 1, 2001; and
- U.S. Patent Application Serial No. 09/532,492, entitled "High Efficiency, High Performance Communications System Employing Multi-Carrier Modulation," filed March 30, 2000.

These applications are all assigned to the assignee of the present application and incorporated herein by reference. Application Serial No. 09/776,075 describes a coding scheme whereby different rates may be achieved by coding the data with the same base code (e.g., a convolutional or Turbo code) and adjusting the puncturing to achieve the

desired rate. Other coding and modulation schemes may also be used, and this is within the scope of the invention.

Receiver System

[1096] FIG. 6 is a block diagram of a RX MIMO/data processor 160a capable of implementing the successive cancellation receiver processing technique. RX MIMO/data processor 160a is one embodiment of RX MIMO/data processor 160 in FIG. 1. The signals transmitted from N_T transmit antennas are received by each of N_R antennas 152a through 152r and routed to a respective receiver 154. Each receiver 154 conditions (e.g., filters, amplifies, and downconverts) a respective received signal and digitizes the conditioned signal to provide a corresponding stream of data samples.

[1097] For MIMO without OFDM, the data samples are representative of the received symbols. Each receiver 154 would then provide to RX MIMO/data processor 160a a respective received symbol stream, which includes a received symbol for each symbol period.

[1098] For MIMO with OFDM, each receiver 154 further includes a cyclic prefix removal element and an FFT processor (both of which are not shown in FIG. 6 for simplicity). The cyclic prefix removal element removes the cyclic prefix, which has been inserted at the transmitter system for each transmission symbol, to provide a corresponding received OFDM symbol. The FFT processor then transforms each received OFDM symbol to provide a vector of N_F received symbols for the N_F frequency subchannels for that symbol period. N_R received symbol vector streams are then provided by N_R receivers 154 to RX MIMO/data processor 160a.

[1099] For MIMO with OFDM, RX MIMO/data processor 160a may demultiplex the N_R received symbol vector streams into N_F groups of N_R received symbol streams, one group for each frequency subchannel, with each group including N_R streams of received symbols for one frequency subchannel. RX MIMO/data processor 160a may then process each group of N_R received symbol streams in similar manner as for the N_R received symbol streams for MIMO without OFDM. RX MIMO/data processor 160a may also process the received symbols for MIMO with OFDM based on some other ordering scheme, as is known in the art. In any case, RX MIMO/data processor 160a processes the N_R received symbol streams (for MIMO without OFDM) or each group of N_R received symbol streams (for MIMO with OFDM).

[1100] In the embodiment shown in FIG. 6, RX MIMO/data processor 160a includes a number of successive (i.e., cascaded) receiver processing stages 610a through 610n, one stage for each of the transmitted data streams to be recovered. Each receiver processing stage 610 (except for the last stage 610n) includes a spatial processor 620, an RX data processor 630, and an interference canceller 640. The last stage 610n includes only spatial processor 620n and RX data processor 630n.

[1101] For the first stage 610a, spatial processor 620a receives and processes the N_R received symbol streams (denoted as the vector $\underline{\mathbf{y}}^1$) from receivers 154a through 154r based on a particular spatial or space-time receiver processing technique to provide (up to) N_T detected symbol streams (denoted as the vector $\underline{\hat{\mathbf{x}}}^1$). For MIMO with OFDM, the N_R received symbol streams comprise the received symbols for one frequency subchannel. The detected symbol stream corresponding to the lowest data rate, $\hat{x}_1$, is selected and provided to RX data processor 630a. Processor 630a further processes (e.g., demodulates, deinterleaves, and decodes) the detected symbol stream, $\hat{x}_1$, selected for the first stage to provide a decoded data stream. Spatial processor 620a further provides an estimate of the channel response matrix $\underline{\mathbf{H}}$, which is used to perform the spatial or space-time processing for all stages.

[1102] For the first stage 610a, interference canceller 640a also receives the N_R received symbol streams from receivers 154 (i.e., the vector $\underline{\mathbf{y}}^1$). Interference canceller 640a further receives the decoded data stream from RX data processor 630a and performs the processing (e.g., encoding, interleaving, modulation, channel response, and so on) to derive N_R remodulated symbol streams (denoted as the vector $\underline{\mathbf{i}}^1$) that are estimates of the interference components due to the just-recovered data stream. The remodulated symbol streams are then subtracted from the first stage's input symbol streams to derive N_R modified symbol streams (denoted as the vector $\underline{\mathbf{y}}^2$), which include all but the subtracted (i.e., cancelled) interference components. The N_R modified symbol streams are then provided to the next stage.

[1103] For each of the second through last stages 610b through 610n, the spatial processor for that stage receives and processes the N_R modified symbol streams from the interference canceller in the preceding stage to derive the detected symbol streams for that stage. The detected symbol stream corresponding to the lowest data rate at that

stage is selected and processed by the RX data processor to provide the decoded data stream for that stage. For each of the second through second-to-last stages, the interference canceller in that stage receives the N_R modified symbol streams from the interference canceller in the preceding stage and the decoded data stream from the RX data processor within the same stage, derives N_R remodulated symbol streams, and provides N_R modified symbol streams for the next stage.

[1104] The successive cancellation receiver processing technique is described in further detail in the aforementioned U.S. Patent Application Serial Nos. 09/993,087 and 09/854,235.

[1105] The spatial processor 620 in each stage implements a particular spatial or space-time receiver processing technique. The specific receiver processing technique to be used is typically dependent on the characteristics of the MIMO channel, which may be characterized as either non-dispersive or dispersive. A non-dispersive MIMO channel experiences flat fading (i.e., approximately equal amount of attenuation across the system bandwidth), and a dispersive MIMO channel experiences frequency-selective fading (e.g., different amounts of attenuation across the system bandwidth).

[1106] For a non-dispersive MIMO channel, spatial receiver processing techniques may be used to process the received signals to provide the detected symbol streams. These spatial receiver processing techniques include a channel correlation matrix inversion (CCMI) technique (which is also referred to as a zero-forcing technique) and a minimum mean square error (MMSE) technique. Other spatial receiver processing techniques may also be used and are within the scope of the invention.

[1107] For a dispersive MIMO channel, time dispersion in the channel introduces inter-symbol interference (ISI). To improve performance, a receiver attempting to recover a particular transmitted data stream would need to ameliorate both the interference (or "crosstalk") from the other transmitted data streams as well as the ISI from all data streams. To combat both crosstalk and ISI, space-time receiver processing techniques may be used to process the received signals to provide the detected symbol streams. These space-time receiver processing techniques include a MMSE linear equalizer (MMSE-LE), a decision feedback equalizer (DFE), a maximum-likelihood sequence estimator (MLSE), and so on.

[1108] The CCMI, MMSE, MMSE-LE, and DFE techniques are described in detail in the aforementioned U.S. Patent Application Serial Nos. 09/993,087, 09/854,235, 09/826,481, and 09/956,44.

[1109] The data rate determination and data transmission techniques described herein may be implemented by various means. For example, these techniques may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the elements used to determinate data rates at the transmitter and the data transmission at the transmitter/receiver may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[1110] For a software implementation, certain aspects of the data rate determination and the processing at the transmitter/receiver may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memory 132 in FIG. 1) and executed by a processor (e.g., controller 130). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as is known in the art.

[1111] Headings are included herein for reference and to aid in locating certain sections. These headings are not intended to limit the scope of the concepts described therein under, and these concepts may have applicability in other sections throughout the entire specification.

[1112] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[1113] **WHAT IS CLAIMED IS:**

CLAIMS

1. A method for determining data rates for a plurality of data streams to be transmitted via a plurality of transmission channels in a multi-channel communication system, comprising:

determining a required signal-to-noise-and-interference ratio (SNR) for each of a plurality of data rates to be used for the plurality of data streams, wherein at least two of the data rates are unequal;

determining an effective SNR for each of the plurality of data streams based in part on successive interference cancellation processing at a receiver to recover the plurality of data streams;

comparing the required SNR for each data stream against the effective SNR for the data stream; and

determining whether or not the plurality of data rates are supported based on results of the comparing.

2. The method of claim 1, wherein the plurality of data streams are transmitted over a plurality of transmit antennas in a multiple-input multiple-output (MIMO) communication system.

3. The method of claim 2, wherein each data stream is transmitted over a respective transmit antenna, and wherein the effective SNR for each data stream is determined based on full transmit power being used for the data stream.

4. The method of claim 1, wherein the effective SNR for each data stream is further determined based on a received SNR indicative of an operating condition of the plurality of transmission channels.

5. The method of claim 4, wherein the received SNR is determined based on the required SNR for one of the plurality of data streams.

6. The method of claim 4, wherein the received SNR is specified for the communication system.

7. The method of claim 4, wherein the received SNR is estimated at the receiver.

8. The method of claim 4, wherein the successive interference cancellation processing recovers one data stream at each stage, and wherein the effective SNR for each recovered data stream is estimated as

$$\mathrm{SNR}_{\mathrm{eff}}(k) = \left(\frac{N_R - N_T + k}{N_T N_R} \right) \mathrm{SNR}_{\mathrm{rx}} \ , \qquad \text{Eq (9)}$$

where $\mathrm{SNR}_{\mathrm{eff}}(k)$ is the effective SNR for the data stream recovered in stage k,

$\mathrm{SNR}_{\mathrm{rx}}$ is the received SNR,

N_T is the number of transmit antennas used for data transmission, and

N_R is the number of receive antennas.

9. The method of claim 4, further comprising:

evaluating a plurality of sets of data rates; and

selecting a rate set associated with a minimum received SNR for use for the plurality of data streams.

10. The method of claim 9, wherein the data rates in each rate set are selected to achieve a specified overall spectral efficiency.

11. The method of claim 1, wherein the required SNR for each data rate is determined based on a look-up table.

12. The method of claim 1, wherein the plurality of data rates are deemed to be supported if the required SNR for each data rate is less than or equal to the effective SNR for the data rate.

13. The method of claim 1, wherein the communication system implements orthogonal frequency division multiplexing (OFDM).

14. A method for determining data rates for a plurality of data streams to be transmitted over a plurality of transmit antennas in a multiple-input multiple-output (MIMO) communication system, comprising:

determining an operating signal-to-noise-and-interference ratio (SNR) indicative of an operating condition of the MIMO system;

determining a required SNR for each of a plurality of data rates to be used for the plurality of data streams, wherein at least two of the data rates are unequal and wherein the plurality of data rates are selected to achieve a specified overall spectral efficiency;

determining an effective SNR for each of the plurality of data streams based on the operating SNR and successive interference cancellation processing technique at a receiver to recover the plurality of data streams;

comparing the required SNR for each data stream against the effective SNR for the data stream; and

determining whether or not the plurality of data rates are supported based on results of the comparing.

15. A method for determining data rates for a plurality of data streams to be transmitted via a plurality of transmission channels in a multi-channel communication system, comprising:

determining a received SNR indicative of an operating condition of the plurality of transmission channels;

determining an effective SNR for each of the plurality of data streams based on the received SNR and successive interference cancellation processing at a receiver to recover the plurality of data streams; and

determining a data rate for each data stream based on the effective SNR for the data stream, wherein at least two of the data rates are unequal.

16. The method of claim 15, wherein the data rate for each data stream is determined such that a required SNR for the data stream is less than or equal to the effective SNR for the data stream.

17. The method of claim 15, wherein the received SNR is specified for the communication system.

18. The method of claim 15, wherein each data stream is transmitted over a respective transmit antenna in a multiple-input multiple-output (MIMO) communication system.

19. A memory communicatively coupled to a digital signal processing device (DSPD) capable of interpreting digital information to:

determine a required signal-to-noise-and-interference ratio (SNR) for each of a plurality of data rates to be used for a plurality of data streams to be transmitted via a plurality of transmission channels in a multi-channel communication system, wherein at least two of the data rates are unequal;

determine an effective SNR for each of the plurality of data streams based in part on successive interference cancellation processing at a receiver to recover the plurality of data streams;

compare the required SNR for each data stream against the effective SNR for the data stream; and

determine whether or not the plurality of data rates are supported based on results of the comparison.

20. An apparatus in a multi-channel communication system, comprising:

means for determining a required signal-to-noise-and-interference ratio (SNR) for each of a plurality of data rates to be used for a plurality of data streams to be transmitted via a plurality of transmission channels, wherein at least two of the data rates are unequal;

means for determining an effective SNR for each of the plurality of data streams based in part on successive interference cancellation processing at a receiver to recover the plurality of data streams;

means for comparing the required SNR for each data stream against the effective SNR for the data stream; and

means for determining whether or not the plurality of data rates are supported based on results of the comparing.

21. The apparatus of claim 20, further comprising:

means for evaluating a plurality of sets of data rates; and

means for selecting a rate set associated with a minimum received SNR for use for the plurality of data streams.

22. The apparatus of claim 20, wherein the multi-channel communication system is a multiple-input multiple-output (MIMO) communication system.

23. The apparatus of claim 22, wherein the MIMO system implements orthogonal frequency division multiplexing (OFDM).

24. A base station comprising the apparatus of claim 20.

25. A wireless terminal comprising the apparatus of claim 20.

26. A transmitter unit in a multiple-input multiple-output (MIMO) communication system, comprising:

a controller operative to determine a plurality of data rates for a plurality of data streams to be transmitted over a plurality of transmit antennas by

determining a required signal-to-noise-and-interference ratio (SNR) for each of the plurality of data rates, wherein at least two of the data rates are unequal,

determining an effective SNR for each of the plurality of data streams based in part on successive interference cancellation processing technique at a receiver to recover the plurality of data streams,

comparing the required SNR for each data stream against the effective SNR for the data stream, and

determining whether or not the plurality of data rates are supported based on results of the comparing;

a transmit (TX) data processor operative to process each data stream with the determined data rate to provide a respective symbol stream; and

one or more transmitters operative to process a plurality of symbol streams for the plurality of data streams to provide a plurality of modulated signals suitable for transmission over the plurality of transmit antennas.

27. The transmitter unit of claim 26, wherein the controller is further operative to determine the data rates for the plurality of data streams by

evaluating a plurality of sets of data rates, and

selecting a rate set associated with a minimum received SNR.

28. A base station comprising the transmitter unit of claim 26.

29. A wireless terminal comprising the transmitter unit of claim 26.

30. A transmitter apparatus in a multiple-input multiple-output (MIMO) communication system, comprising:

means for determining a required signal-to-noise-and-interference ratio (SNR) for each of a plurality of data rates to be used for a plurality of data streams to be transmitted over a plurality of transmit antennas in the MIMO system, wherein at least two of the data rates are unequal;

means for determining an effective SNR for each of the plurality of data streams based in part on successive interference cancellation processing at a receiver to recover the plurality of data streams;

means for comparing the required SNR for each data stream against the effective SNR for the data stream;

means for determining whether or not the plurality of data rates are supported based on results of the comparison;

means for processing each data stream to provide a respective symbol stream; and

means for processing a plurality of symbol streams for the plurality of data streams to provide a plurality of modulated signals suitable for transmission over the plurality of transmit antennas.

31. A receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

a receive (RX) MIMO processor operative to receive and process a plurality of received symbol streams using successive interference cancellation processing to provide a plurality of detected symbol streams for a plurality of transmitted data streams, one detected data stream for each stage of the successive interference cancellation processing; and

a RX data processor operative to process each detected symbol stream to provide a corresponding decoded data stream, and

wherein data rates for the plurality of transmitted data streams are determined by determining a received signal-to-noise-and-interference ratio (SNR) indicative of an operating condition of the communication system, determining an effective SNR for each of the plurality of data streams based on the received SNR and the successive interference cancellation processing, and determining the data rate for each data stream based on the effective SNR, and wherein at least two of the data rates are unequal.

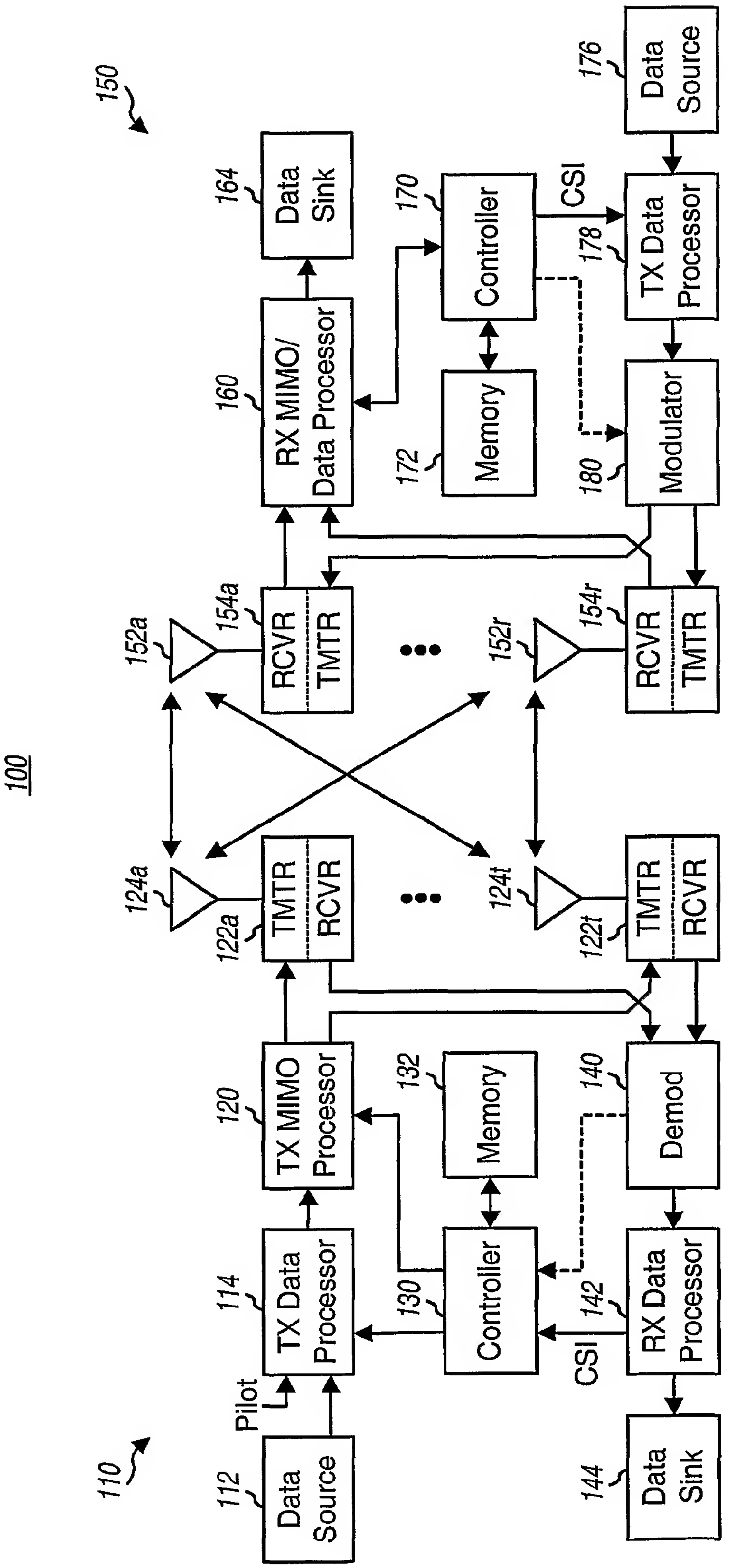

100
110
112
Data Source
Pilot
114
TX Data Processor
120
TX MIMO Processor
130
Controller
132
Memory
122a
TMTR
RCVR
124a
122t
124t
140
Demod
142
RX Data Processor
CSI
144
Data Sink
150
152a
154a
RCVR
TMTR
152r
154r
160
RX MIMO/ Data Processor
164
Data Sink
170
Controller
172
Memory
176
Data Source
178
TX Data Processor
180
Modulator
CSI

FIG. 1

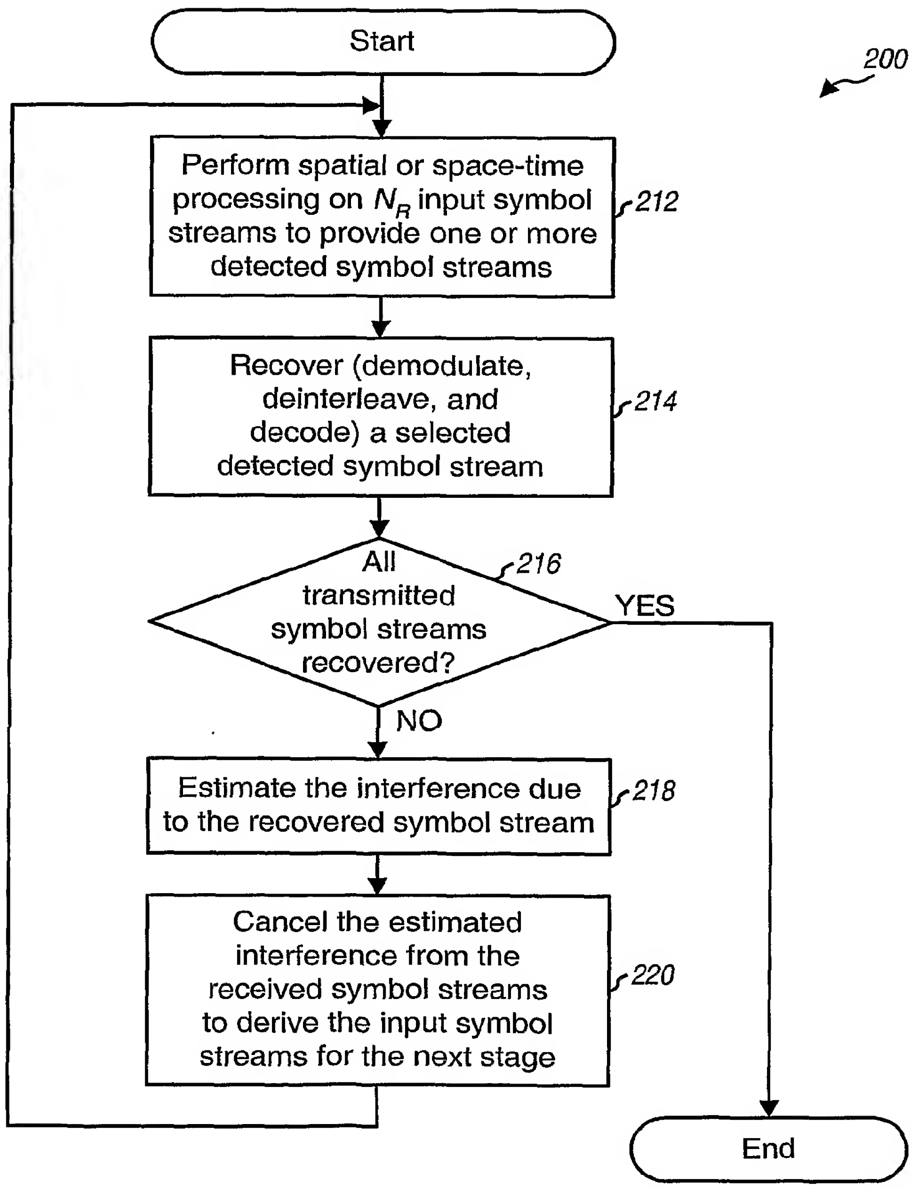

200
Start
Perform spatial or space-time processing on N_R input symbol streams to provide one or more detected symbol streams
212
Recover (demodulate, deinterleave, and decode) a selected detected symbol stream
214
All transmitted symbol streams recovered?
216
YES
NO
Estimate the interference due to the recovered symbol stream
218
Cancel the estimated interference from the received symbol streams to derive the input symbol streams for the next stage
220
End

FIG. 2

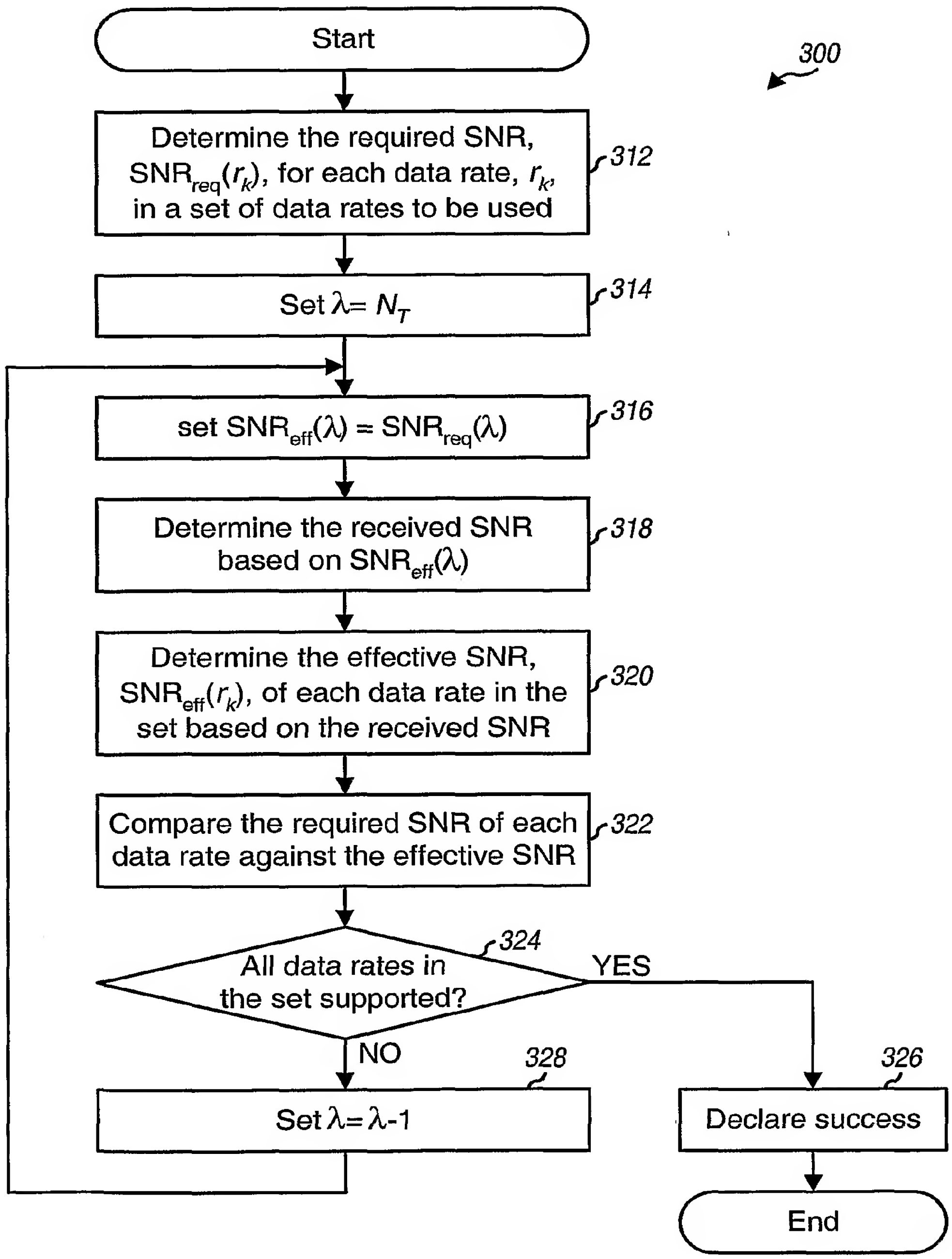
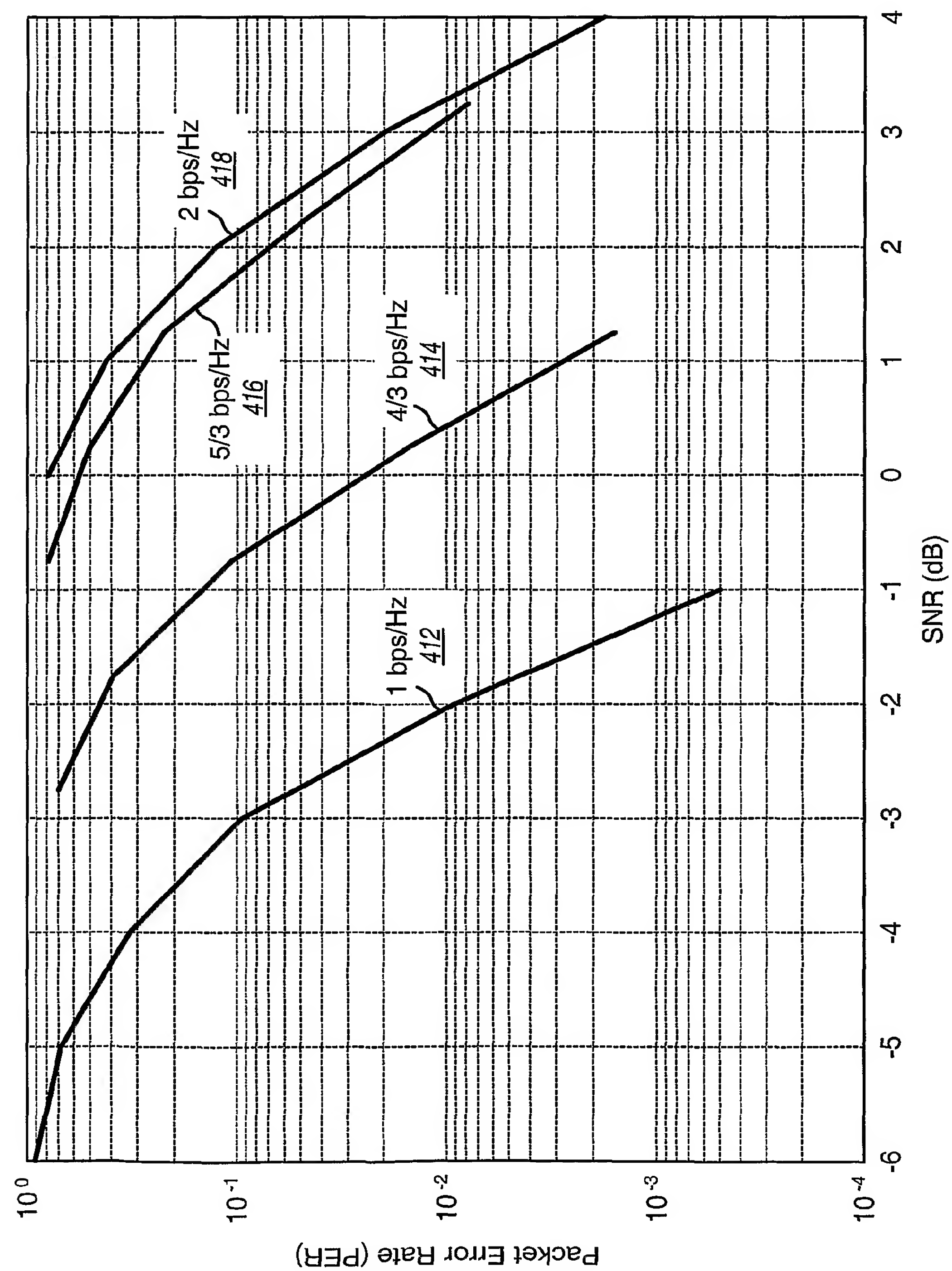

300
Start
Determine the required SNR, SNRreq(rk), for each data rate, rk, in a set of data rates to be used
312
Set λ= NT
314
set SNReff(λ) = SNRreq(λ)
316
Determine the received SNR based on SNReff(λ)
318
Determine the effective SNR, SNReff(rk), of each data rate in the set based on the received SNR
320
Compare the required SNR of each data rate against the effective SNR
322
All data rates in the set supported?
324
YES
NO
Set λ=λ-1
328
Declare success
326
End

FIG. 3



Packet Error Rate (PER)
10^0
10^-1
10^-2
10^-3
10^-4
SNR (dB)
-6
-5
-4
-3
-2
-1
0
1
2
3
4
1 bps/Hz
412
4/3 bps/Hz
414
5/3 bps/Hz
416
2 bps/Hz
418

FIG. 4

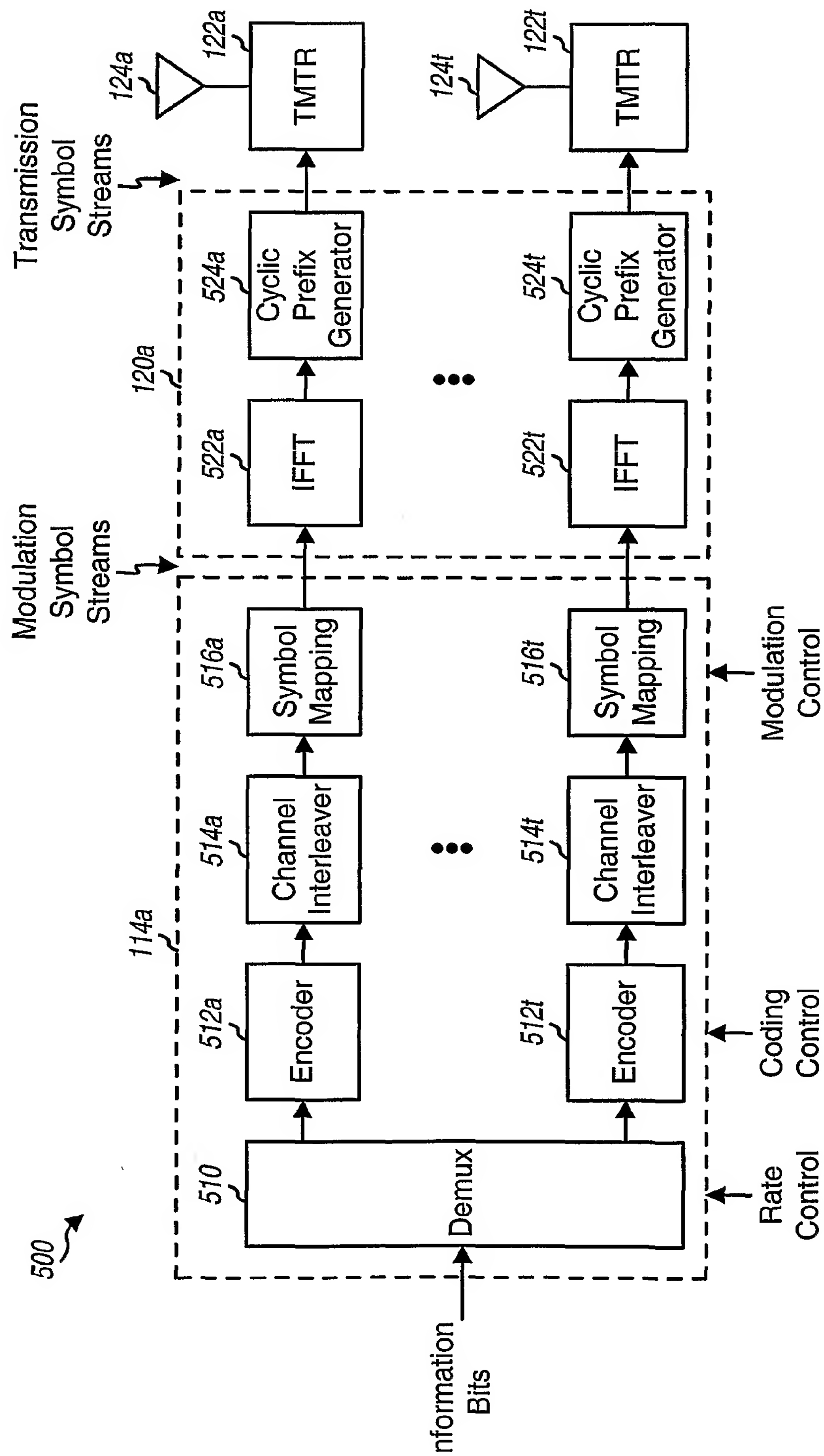

500
Information Bits
114a
510
Demux
512a
Encoder
514a
Channel Interleaver
516a
Symbol Mapping
512t
Encoder
514t
Channel Interleaver
516t
Symbol Mapping
Rate Control
Coding Control
Modulation Control
Modulation Symbol Streams
120a
522a
IFFT
524a
Cyclic Prefix Generator
522t
IFFT
524t
Cyclic Prefix Generator
Transmission Symbol Streams
122a
TMTR
124a
122t
TMTR
124t

FIG. 5

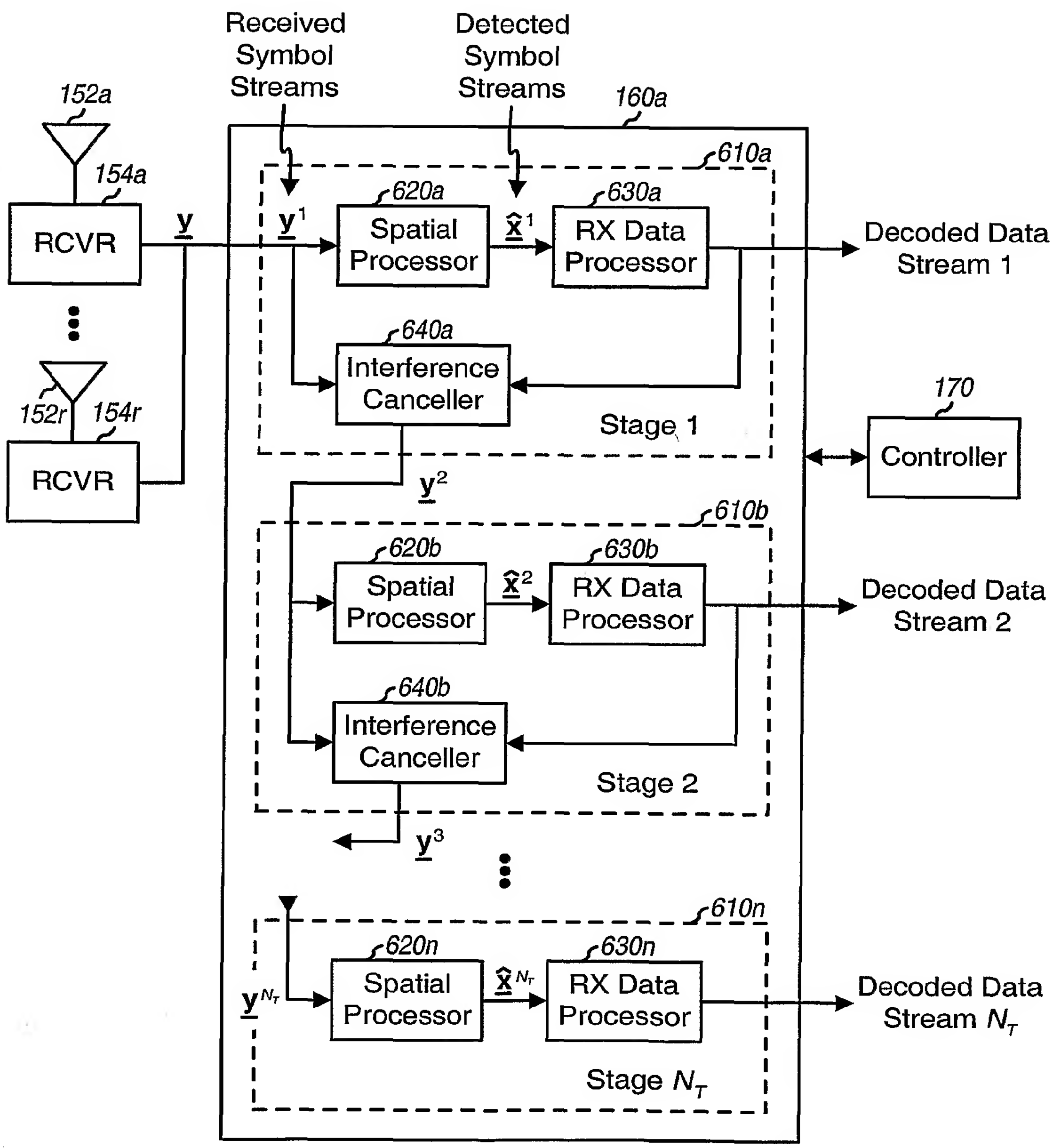

Received
Symbol
Streams
Detected
Symbol
Streams
152a
154a
RCVR
y
160a
610a
620a
630a
y1
Spatial
Processor
x̂1
RX Data
Processor
Decoded Data
Stream 1
640a
Interference
Canceller
Stage 1
152r
154r
RCVR
170
Controller
y2
610b
620b
630b
Spatial
Processor
x̂2
RX Data
Processor
Decoded Data
Stream 2
640b
Interference
Canceller
Stage 2
y3
610n
620n
630n
yNT
Spatial
Processor
x̂NT
RX Data
Processor
Decoded Data
Stream NT
Stage NT

FIG. 6

INTERNATIONAL SEARCH REPORT	International application No. PCT/US03/06326

A. CLASSIFICATION OF SUBJECT MATTER

IPC(7) :HO4B 1/69; HO4J 11/00
US CL :375/130; 370/208
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 375/130, 220, 284, 285; 370/208, 286, 289, 278

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

EAST

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	US 6,154,484 A (LEE et al) 28 November 2000 see figs 1-5 and col.2, lines 5-15 and col.4, lines 60-67 and col.7, lines 35-50 and col.14, lines 11-20 and col.22, lines 55-57	1-31
Y	US 6,141,317 A (MARCHOK et al)31 October 2000 see fig.2 and col.22, lines 44-67 and col.27, lines 1-67 and col.28, lines 1-67	1-31

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
17 APRIL 2003	12 MAY 2003
Name and mailing address of the ISA/US Commissioner of Patents and Trademarks Box PCT Washington, D.C. 20231 Facsimile No. (703) 305-3230	Authorized officer BAYARD, EMMANUEL Telephone No. (703) 308-9573

Form PCT/ISA/210 (second sheet) (July 1998)★

Box I Observations where certain claims were found unsearchable (Continuation of item 1 of first sheet)

This international report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. ☐ Claims Nos.:
because they relate to subject matter not required to be searched by this Authority, namely:

2. ☐ Claims Nos.:
because they relate to parts of the international application that do not comply with the prescribed requirements to such an extent that no meaningful international search can be carried out, specifically:

3. ☐ Claims Nos.:
because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a).

Box II Observations where unity of invention is lacking (Continuation of item 2 of first sheet)

This International Searching Authority found multiple inventions in this international application, as follows:

1. ☐ As all required additional search fees were timely paid by the applicant, this international search report covers all searchable claims.

2. ☐ As all searchable claims could be searched without effort justifying an additional fee, this Authority did not invite payment of any additional fee.

3. ☐ As only some of the required additional search fees were timely paid by the applicant, this international search report covers only those claims for which fees were paid, specifically claims Nos.:

4. ☐ No required additional search fees were timely paid by the applicant. Consequently, this international search report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

Remark on Protest

☐ The additional search fees were accompanied by the applicant's protest.

☐ No protest accompanied the payment of additional search fees.

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

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International Bureau

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12 May 2005 (12.05.2005)

PCT

(10) International Publication Number
WO 2005/043855 A1

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(26) Publication Language: English

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10/778,570	13 February 2004 (13.02.2004)	US

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(75) **Inventor/Applicant** *(for US only)*: **KADOUS, Tamer** [EG/US]; 5385 Toscana Way, #316, San Diego, California 92122 (US).

(74) **Agents: WADSWORTH, Philip, R.** et al.; 5775 Morehouse Drive, San Diego, California 92121 (US).

(81) **Designated States** *(unless otherwise indicated, for every kind of national protection available)*: AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BW, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NA, NI, NO, NZ, OM, PG, PH, PL, PT, RO, RU, SC, SD, SE, SG, SK, SL, SY, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, YU, ZA, ZM, ZW.

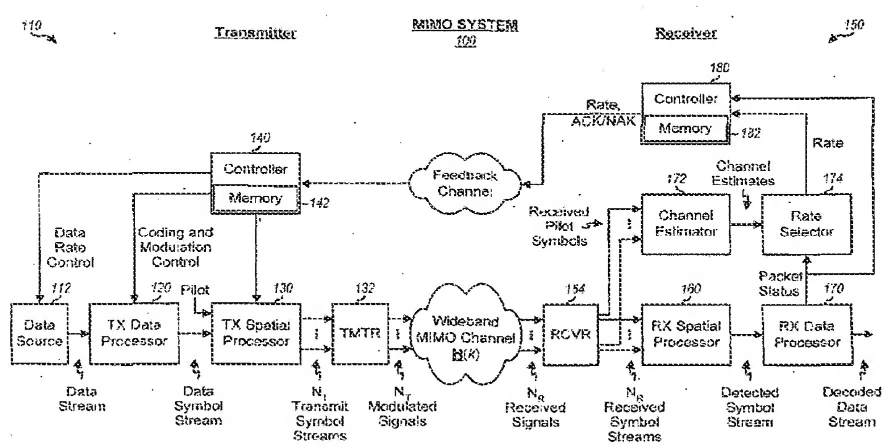
(84) **Designated States** *(unless otherwise indicated, for every kind of regional protection available)*: ARIPO (BW, GH, GM, KE, LS, MW, MZ, NA, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HU, IE, IT, LU, MC, NL, PL, PT, RO, SE, SI, SK, TR), OAPI (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Declarations under Rule 4.17:

— *as to applicant's entitlement to apply for and be granted a patent (Rule 4.17(ii)) for the following designations AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BW, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE,*

[Continued on next page]

(54) **Title:** RATE SELECTION FOR A MULTI-CARRIER MIMO SYSTEM



(57) **Abstract:** To select a rate for data transmission in a multi-carrier MIMO system with a multipath MIMO channel, a post-detection SNR for each subband k of each spatial channel is initially determined and used to derive a constrained spectral efficiency based on a constrained spectral efficiency function of SNR and modulation scheme M. An average constrained spectral efficiency for all subbands of all spatial channels used for data transmission is next determined based on the constrained spectral efficiencies for the individual subbands/spatial channels. An equivalent SNR needed by an equivalent system with an AWGN channel to support a data rate of is determined based on an inverse constrained spectral efficiency function . A rate is selected for the multi-carrier MIMO system based on the equivalent SNR. The selected rate is the highest rate among all supported rates with a required SNR less than or equal to the equivalent SNR.

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RATE SELECTION FOR A MULTI-CARRIER MIMO SYSTEM

CROSS-REFERENCE TO RELATED APPLICATION

[0001] This application claims the benefit of U.S. Provisional Patent Application Serial No. 60/514,402, filed October 24, 2003, which is incorporated herein by reference in their entirety.

I. Field

[0002] The present invention relates generally to communication, and more specifically to techniques for performing rate selection for data transmission in a multi-carrier multiple-input multiple-output (MIMO) communication system.

II. Background

[0003] A MIMO system employs multiple (N_T) transmit antennas at a transmitter and multiple (N_R) receive antennas at a receiver for data transmission. A MIMO channel formed by the N_T transmit antennas and N_R receive antennas may be decomposed into N_S spatial channels, where $N_S \leq \min\{N_T, N_R\}$. The N_S spatial channels may be used to transmit data in parallel to achieve higher throughput and/or redundantly to achieve greater reliability.

[0004] Orthogonal frequency division multiplexing (OFDM) is a multi-carrier modulation scheme that effectively partitions the overall system bandwidth into multiple (N_F) orthogonal subbands. These subbands are also referred to as tones, subcarriers, bins, and frequency channels. With OFDM, each subband is associated with a respective subcarrier that may be modulated with data.

[0005] For a MIMO system that utilizes OFDM (i.e., a MIMO-OFDM system), N_F subbands are available on each of the N_S spatial channels for data transmission. The N_F subbands of each spatial channel may experience different channel conditions (e.g., different fading, multipath, and interference effects) and may achieve different channel gains and signal-to-noise-and-interference ratios (SNRs). Depending on the multipath profile of the MIMO channel, the channel gains and SNRs may vary widely across the

N_F subbands of each spatial channel and may further vary widely among the N_S spatial channels.

[0006] For the MIMO-OFDM system, one modulation symbol may be transmitted on each subband of each spatial channel, and up to $N_F \cdot N_S$ modulation symbols may be transmitted simultaneously in each OFDM symbol period. Each transmitted modulation symbol is distorted by the channel gain for the subband of the spatial channel via which the symbol is transmitted and further degraded by channel noise and interference. For a multipath MIMO channel, which is a MIMO channel with a frequency response that is not flat, the number of information bits that may be reliably transmitted on each subband of each spatial channel may vary from subband to subband and from spatial channel to spatial channel. The different transmission capabilities of the different subbands and spatial channels plus the time-variant nature of the MIMO channel make it challenging to ascertain the true transmission capacity of the MIMO-OFDM system.

[0007] There is therefore a need in the art for techniques to accurately determine the transmission capacity of the MIMO-OFDM system for efficient data transmission.

SUMMARY

[0008] Techniques for performing rate selection in a multi-carrier MIMO system (e.g., a MIMO-OFDM system) with a multipath MIMO channel are described herein. In an embodiment, a post-detection SNR, $SNR_\ell(k)$, for each subband k of each spatial channel ℓ used for data transmission is initially determined for a "theoretical" multi-carrier MIMO system that is capable of achieving capacity of the MIMO channel. The post-detection SNR is the SNR after spatial processing or detection at a receiver. The theoretical system has no implementation losses. A constrained spectral efficiency $S_\ell(k)$ for each subband of each spatial channel is then determined based on its post-detection SNR, a modulation scheme M, and a constrained spectral efficiency function $f_{siso}(SNR_\ell(k),\ M)$. An average constrained spectral efficiency S_{avg} for all subbands of all spatial channels used for data transmission is next determined based on the constrained spectral efficiencies for the individual subbands of the spatial channels.

[0009] An equivalent system with an additive white Gaussian noise (AWGN) channel needs an SNR of SNR_{equiv} to achieve a constrained spectral efficiency of S_{avg} with modulation scheme M. An AWGN channel is a channel with a flat frequency response.

The equivalent system also has no implementation losses. The equivalent SNR may be determined based on an inverse constrained spectral efficiency function $f_{siso}^{-1}(S_{avg}, M)$. A rate R is then selected for data transmission in the multi-carrier MIMO system based on the equivalent SNR. The multi-carrier MIMO system may support a specific set of rates, and the required SNRs for these rates may be determined and stored in a look-up table. The selected rate is the highest rate among the supported rates with a required SNR that is less than or equal to the equivalent SNR. A back-off factor may be computed to account for error in the rate prediction, system losses, and so on. The rate R may then be selected in a manner to account for the back-off factor, as described below.

[0010] Various aspects and embodiments of the invention are described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[0011] The features and nature of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[0012] FIG. 1 shows a transmitter and a receiver in a MIMO-OFDM system;

[0013] FIG. 2 illustrates rate selection for the MIMO-OFDM system;

[0014] FIG. 3 shows a process for performing rate selection for a MIMO-OFDM system with a multipath MIMO channel;

[0015] FIG. 4A illustrates constrained spectral efficiencies for N_T spatial channels in the MIMO-OFDM system with the multipath MIMO channel;

[0016] FIG. 4B illustrates constrained spectral efficiency for an equivalent system with an AWGN channel;

[0017] FIG. 5 shows a block diagram of the transmitter;

[0018] FIG. 6 shows a block diagram of the receiver; and

[0019] FIG. 7 shows a receive (RX) spatial processor and an RX data processor that implement iterative detection and decoding (IDD).

DETAILED DESCRIPTION

[0020] The word "exemplary" is used herein to mean "serving as an example, instance, or illustration." Any embodiment or design described herein as "exemplary" is not

necessarily to be construed as preferred or advantageous over other embodiments or designs.

[0021] The rate selection techniques described herein may be used for various types of multi-carrier MIMO system. For clarity, these techniques are specifically described for a MIMO-OFDM system.

[0022] **FIG. 1** shows a block diagram of a transmitter 110 and a receiver 150 in a MIMO-OFDM system 100. At transmitter 110, a transmit (TX) data processor 120 receives packets of data from a data source 112. TX data processor 120 encodes, interleaves, and modulates each data packet in accordance with a rate selected for that packet to obtain a corresponding block of data symbols. As used herein, a data symbol is a modulation symbol for data, and a pilot symbol is a modulation symbol for pilot, which is known *a priori* by both the transmitter and receiver. The selected rate for each data packet may indicate the data rate, coding scheme or code rate, modulation scheme, packet size, and so on for that packet, which are indicated by various controls provided by a controller 140.

[0023] A TX spatial processor 130 receives and spatially processes each data symbol block for transmission on the N_F subbands of the N_T transmit antennas. TX spatial processor 130 further multiplexes in pilot symbols and provides N_T streams of transmit symbols to a transmitter unit (TMTR) 132. Each transmit symbol may be for a data symbol or a pilot symbol. Transmitter unit 132 performs OFDM modulation on the N_T transmit symbol streams to obtain N_T OFDM symbol streams and further processes these OFDM symbol streams to generate N_T modulated signals. Each modulated signal is transmitted from a respective transmit antenna (not shown in FIG. 1) and via a MIMO channel to receiver 150. The MIMO channel distorts the N_T transmitted signals with a MIMO channel response and further degrades the transmitted signals with noise and possibly interference from other transmitters.

[0024] At receiver 150, the N_T transmitted signals are received by each of N_R receive antennas (not shown in FIG. 1), and the N_R received signals from the N_R receive antennas are provided to a receiver unit (RCVR) 154. Receiver unit 154 conditions and digitizes each received signal to obtain a corresponding stream of samples and further performs OFDM demodulation on each sample stream to obtain a stream of received symbols. Receiver unit 154 provides N_R received symbol streams (for data) to an RX spatial processor 160 and received pilot symbols (for pilot) to a channel estimator 172. RX spatial processor 160 spatially processes or detects the N_R received symbol streams

to obtain detected symbols, which are estimates of the data symbols transmitted by transmitter 110.

[0025] An RX data processor 170 receives, demodulates, deinterleaves, and decodes each block of detected symbols in accordance with its selected rate to obtain a corresponding decoded packet, which is an estimate of the data packet sent by transmitter 110. RX data processor 170 also provides the status of each decoded packet, which indicates whether the packet is decoded correctly or in error.

[0026] Channel estimator 172 processes the received pilot symbols and/or received data symbols to obtain channel estimates for the MIMO channel. The channel estimates may include channel gain estimates, SNR estimates, and so on. A rate selector 174 receives the channel estimates and selects a suitable rate for data transmission to receiver 150. A controller 180 receives the selected rate from rate selector 174 and the packet status from RX data processor 170 and assembles feedback information for transmitter 110. The feedback information may include the selected rate, acknowledgments (ACKs) or negative acknowledgments (NAKs) for current and/or prior data packets, and so on. The feedback information is processed and transmitted via a feedback channel to transmitter 110.

[0027] At transmitter 110, the signal(s) transmitted by receiver 150 are received and processed to recover the feedback information sent by receiver 150. Controller 140 receives the recovered feedback information, uses the selected rate to process subsequent data packets to be sent to receiver 150, and uses the ACKs/NAKs to control retransmission of the current and/or prior data packets.

[0028] Controllers 140 and 180 direct the operation at transmitter 110 and receiver 150, respectively. Memory units 142 and 182 provide storage for program codes and data used by controllers 140 and 180, respectively. Memory units 142 and 182 may be internal to controllers 140 and 180, as shown in FIG. 1, or external to these controllers.

[0029] A major challenge for the MIMO-OFDM system is selecting a suitable rate for data transmission based on channel conditions. The goal of the rate selection is to maximize throughput on the N_S spatial channels while meeting certain quality objectives, which may be quantified by a particular packet error rate (e.g., 1% PER).

[0030] The performance of the MIMO-OFDM system is highly dependent on the accuracy of the rate selection. If the selected rate for data transmission is too conservative, then excessive system resources are expended for the data transmission and channel capacity is underutilized. Conversely, if the selected rate is too aggressive, then the receiver may

decode the data transmission in error and system resources may be expended for retransmission. Rate selection for the MIMO-OFDM system is challenging because of the complexity in estimating the true transmission capability of a multipath MIMO channel.

[0031] A multipath MIMO channel formed by the N_T transmit antennas at transmitter 110 and the N_R receive antennas at receiver 150 may be characterized by a set of N_F channel response matrices $\underline{\mathbf{H}}(k)$, for $k = 1 \ldots N_F$, which may be expressed as:

$$\underline{\mathbf{H}}(k) = \begin{bmatrix} h_{1,1}(k) & h_{1,2}(k) & \cdots & h_{1,N_T}(k) \\ h_{2,1}(k) & h_{2,2}(k) & \cdots & h_{2,N_T}(k) \\ \vdots & \vdots & \ddots & \vdots \\ h_{N_R,1}(k) & h_{N_R,2}(k) & \cdots & h_{N_R,N_T}(k) \end{bmatrix}, \quad \text{for } k = 1 \ldots N_F, \qquad \text{Eq (1)}$$

where entry $h_{i,j}(k)$, for $i = 1 \ldots N_R$, $j = 1 \ldots N_T$, and $k = 1 \ldots N_F$, denotes the complex channel gain between transmit antenna j and receive antenna i for subband k. For simplicity, the following description assumes that each channel response matrix $\underline{\mathbf{H}}(k)$ is full rank and the number of spatial channels is $N_S = N_T \le N_R$. In general, a spatial channel is an effective channel between an element of a data symbol vector $\underline{s}(k)$ at the transmitter and a corresponding element of a detected symbol vector $\hat{\underline{s}}(k)$ at the receiver. The vectors $\underline{s}(k)$ and $\hat{\underline{s}}(k)$ are described below. The N_T spatial channels of the MIMO channel are dependent on the spatial processing (if any) performed at the transmitter and the spatial processing performed at the receiver.

[0032] The multipath MIMO channel has a capacity that can be determined in various manners. As used herein, "capacity" denotes the transmission capability of a channel, and "spectral efficiency" denotes the general concept of "capacity per dimension", where the dimension may be frequency and/or space. Spectral efficiency may be given in units of bits per second per Hertz per spatial channel (bps/Hz/ch) for the MIMO-OFDM system. Spectral efficiency is often specified as being either constrained or unconstrained. An "unconstrained" spectral efficiency is typically defined as the theoretical maximum data rate that may be reliably used for a channel with a given channel response and noise variance. A "constrained" spectral efficiency is further dependent on the specific modulation scheme used for data transmission. The constrained capacity (due to the fact that modulation symbols are restricted to specific

points on a signal constellation) is lower than the unconstrained capacity (which is not confined by any signal constellation).

[0033] **FIG. 2** graphically illustrates a technique for performing rate selection for a MIMO-OFDM system with a multipath MIMO channel. For a given multipath MIMO channel defined by a channel response of $\underline{\mathbf{H}}(k)$, for $k = 1 \ldots N_F$, and a noise variance of N_0, a theoretical MIMO-OFDM system has an average constrained spectral efficiency of S_{avg} with modulation scheme M. As used herein, a "theoretical" system is one without any losses, and a "practical" system is one with implementation losses (e.g., due to hardware imperfections), code loss due to the fact that practical codes do not work at capacity, and any other losses. The theoretical and practical systems both use one or more modulation schemes for data transmission and are defined by constrained spectral efficiencies. The average constrained spectral efficiency S_{avg} may be determined as described below. In general, different modulation schemes may be used for different subbands and/or spatial channels. For simplicity, the following description assumes that the same modulation scheme M is used for all subbands of all spatial channels available for data transmission.

[0034] An equivalent system with an AWGN channel needs an SNR of SNR_{equiv} to achieve a constrained spectral efficiency of S_{avg} with modulation scheme M. This equivalent system also has no losses. The equivalent SNR may be derived as described below.

[0035] A practical MIMO-OFDM system with an AWGN channel requires an SNR of SNR_{req} or better to support rate R, which is associated with modulation scheme M, coding scheme C, and data rate D. The data rate D is given in units of bps/Hertz/ch, which is the same unit used for spectral efficiency. The rate R may be selected as the highest rate supported by the system with a required SNR equal to or less than the equivalent SNR, as described below. The required SNR is dependent on modulation scheme M, coding scheme C, and other system losses. The required SNR may be determined for each supported rate (e.g., based on computer simulation, empirical measurement, or some other means) and stored in a look-up table.

[0036] A practical MIMO-OFDM system with a multipath MIMO channel (e.g., MIMO-OFDM system 100) is deemed to support rate R with modulation scheme M and coding scheme C if the required SNR is less than or equal to the equivalent SNR. As

the rate increases, the required SNR increases for the practical system while the equivalent SNR is approximately constant since it is defined by the channel response $\underline{\mathbf{H}}(k)$ and the noise variance N_0. The maximum rate that may be supported by the practical MIMO-OFDM system with the multipath MIMO channel is thus limited by the channel conditions. Details of the rate selection are described below.

[0037] An ideal system having an unconstrained spectral efficiency can be analyzed and used for rate selection for the practical system having a constrained spectral efficiency. An unconstrained spectral efficiency for each subband of the multipath MIMO channel may be determined based on an unconstrained MIMO spectral efficiency function, as follows:

$$S_{unconst}(k) = \frac{1}{\mathrm{N_T}} \cdot \log_2 \left[\det \left(\underline{\mathbf{I}} + \underline{\mathbf{H}}(k) \cdot \underline{\Gamma}(k) \cdot \underline{\mathbf{H}}^H(k) \right) \right], \quad \text{for } k = 1 \ \ldots \ \mathrm{N_F}, \qquad \text{Eq (2)}$$

where $\det(\underline{\mathbf{M}})$ denotes the determinant of $\underline{\mathbf{M}}$, $\underline{\mathbf{I}}$ is the identity matrix, $S_{unconst}(k)$ is the unconstrained spectral efficiency of $\underline{\mathbf{H}}(k)$, $\underline{\Gamma}(k)$ is a matrix that determines the power used for the transmit antennas, and "^{H}" denotes a conjugate transpose. If the channel response $\underline{\mathbf{H}}(k)$ is only known by the receiver, then $\underline{\Gamma}(k)$ is equal to the identity matrix (i.e., $\underline{\Gamma}(k) = \underline{\mathbf{I}}$).

[0038] For a capacity achieving MIMO-OFDM system, which is a system that can transmit and receive data at the capacity of the MIMO channel assuming that a capacity achieving code is available for use, the unconstrained spectral efficiency for each subband of the MIMO channel may be determined based on an unconstrained SISO spectral efficiency function, as follows:

$$S_{unconst}(k) = \frac{1}{\mathrm{N_T}} \cdot \sum_{\ell=1}^{\mathrm{N_T}} \log_2 \left[1 + \mathrm{SNR}_\ell(k) \right], \quad \text{for } k = 1 \ \ldots \ \mathrm{N_F}, \qquad \text{Eq (3)}$$

where $\mathrm{SNR}_\ell(k)$ is the post-detection SNR for subband k of spatial channel ℓ for the capacity achieving system. The post-detection SNR is the SNR achieved for a detected symbol stream after the receiver spatial processing to remove interference from the other symbol streams. The post-detection SNR in equation (3) may be obtained, for example, by a receiver that uses a successive interference cancellation (SIC) technique with a minimum mean square error (MMSE) detector, as described below. Equations

(2) and (3) indicate that, for the capacity achieving system, the unconstrained spectral efficiency of the MIMO channel is equal to the sum of the unconstrained spectral efficiencies of the N_T single-input single-output (SISO) channels that make up the MIMO channel. Each SISO channel corresponds to a spatial channel of the MIMO channel.

[0039] If a single data rate is used for data transmission on all N_F subbands of all N_T transmit antennas, then this single data rate may be set to the average unconstrained spectral efficiency for the N_F subbands of the MIMO channel, as follows:

$$D_{unconst} = \frac{1}{N_F} \cdot \sum_{k=1}^{N_F} S_{unconst}(k) . \qquad \text{Eq (4)}$$

Substituting the unconstrained SISO spectral efficiency function in equation (3) into equation (4), the single data rate may be expressed as:

$$D_{unconst} = \frac{1}{N_F N_T} \cdot \sum_{k=1}^{N_F} \sum_{\ell=1}^{N_T} \log_2 \left[1 + \text{SNR}_\ell(k) \right] . \qquad \text{Eq (5)}$$

[0040] The data rate $D_{unconst}$ is obtained based on the average unconstrained spectral efficiency and is suitable for the ideal MIMO-OFDM system, which is not restricted to a specific modulation scheme. The practical MIMO-OFDM system uses one or more specific modulation schemes for data transmission and has a constrained spectral efficiency that is less than the unconstrained capacity. The data rate $D_{unconst}$ derived based on equation (5) is an optimistic data rate for the practical MIMO-OFDM system. A more accurate data rate may be obtained for the practical MIMO-OFDM system based on a constrained capacity function, instead of an unconstrained capacity function, as described below.

[0041] **FIG. 3** shows a process 300 for performing rate selection for a practical MIMO-OFDM system with a multipath MIMO channel. Process 300 may be performed by rate selector 174 or some other processing unit at the receiver. Initially, an average constrained spectral efficiency S_{avg} for the MIMO channel is determined (block 310). This may be achieved in several ways.

[0042] If a constrained MIMO spectral efficiency function $f_{mimo}(\underline{\mathbf{H}}(k),\ M)$ is available, then the constrained spectral efficiency for each subband of the MIMO channel may be computed based on this function (block 312), as follows:

$$S_{mimo}(k)=\frac{1}{\mathrm{N}_{\mathrm{T}}}\cdot f_{mimo}(\underline{\mathbf{H}}(k),\ M)\ ,\quad \text{for } k=1\ \ldots\ \mathrm{N}_{\mathrm{F}}. \qquad \text{Eq (6)}$$

The average constrained spectral efficiency S_{avg} for all subbands of the MIMO channel may then be computed (block 314), as follows:

$$S_{avg}=\frac{1}{\mathrm{N}_{\mathrm{F}}}\cdot\sum_{k=1}^{\mathrm{N}_{\mathrm{F}}}S_{mimo}(k)\ . \qquad \text{Eq (7)}$$

[0043] The constrained MIMO spectral efficiency function $f_{mimo}(\underline{\mathbf{H}}(k),\ M)$ is likely to be a complex equation with no closed form solution or may not even be available. In this case, the MIMO channel may be decomposed into $\mathrm{N_T}$ SISO channels, and the average constrained spectral efficiency S_{avg} for the MIMO channel may be determined based on the constrained spectral efficiencies of the individual SISO channels. Since the unconstrained spectral efficiency of the MIMO channel is equal to the sum of the unconstrained spectral efficiencies of the $\mathrm{N_T}$ SISO channels for a capacity achieving system, as described above, the constrained spectral efficiency of the MIMO channel can be assumed to be equal to the sum of the constrained spectral efficiencies of the $\mathrm{N_T}$ SISO channels for the capacity achieving system.

[0044] To compute S_{avg}, the post-detection SNR $SNR_{\ell}(k)$ for each subband k of each spatial channel ℓ may be determined for the capacity achieving system, as described below (block 322). The constrained spectral efficiency $S_{\ell}(k)$ for each subband of each spatial channel is then determined based on a constrained SISO spectral efficiency function $f_{siso}(SNR_{\ell}(k),\ M)$ (block 324), as follows:

$$S_{\ell}(k)=f_{siso}(SNR_{\ell}(k),\ M)\ ,\quad \text{for } k=1\ \ldots\ \mathrm{N}_{\mathrm{F}} \text{ and } \ell=1\ \ldots\ \mathrm{N}_{\mathrm{T}}. \qquad \text{Eq (8)}$$

The constrained SISO spectral efficiency function $f_{siso}(SNR_{\ell}(k),\ M)$ may be defined as:

$$f_{siso}(SNR_\ell(k),\ M) = B - \frac{1}{2^B}\sum_{i=1}^{2^B} E\left[\log_2 \sum_{j=1}^{2^B} \exp\left(-\mathrm{SNR}_\ell(k)\cdot(|a_i - a_j|^2 + 2\,\mathrm{Re}\{\eta^*(a_i - a_j)\})\right)\right], \qquad \text{Eq (9)}$$

where B is the number of bits for each modulation symbol for modulation scheme M;

a_i and a_j are signal points in the 2^B-ary constellation for modulation scheme M;

η is a complex Gaussian random variable with zero mean and a variance of $1/\mathrm{SNR}_\ell(k)$; and

$E[\cdot]$ is an expectation operation taken with respect to η in equation (9).

Modulation scheme M is associated with a 2^B-ary constellation (e.g., 2^B-ary QAM) that contains 2^B signal points. Each signal point in the constellation is labeled with a different B-bit value.

[0045] The constrained SISO spectral efficiency function shown in equation (9) does not have a closed form solution. This function may be numerically solved for various SNR values for each modulation scheme, and the results may be stored in a look-up table. Thereafter, the constrained SISO spectral efficiency function may be evaluated by accessing the look-up table with the modulation scheme M and the post-detection SNR $SNR_\ell(k)$.

[0046] The average constrained spectral efficiency S_{avg} for all subbands of all spatial channels may then be computed (block 326), as follows:

$$S_{avg} = \frac{1}{\mathrm{N_F}\cdot\mathrm{N_T}}\cdot\sum_{k=1}^{\mathrm{N_F}}\sum_{\ell=1}^{\mathrm{N_T}} S_\ell(k) \ . \qquad \text{Eq (10)}$$

[0047] The average constrained spectral efficiency S_{avg} may be computed for a practical MIMO-OFDM system with a multipath MIMO channel in various manners. Two exemplary methods are described above. Other methods may also be used.

[0048] An equivalent system with an AWGN channel would require an SNR of SNR_{equiv} to achieve a constrained spectral efficiency of S_{avg} with modulation scheme M. The equivalent SNR may be determined based on an inverse constrained SISO spectral

efficiency function $f_{siso}^{-1}(S_{avg}, M)$ (block 330). The constrained SISO spectral efficiency function $f_{siso}(x)$ takes two inputs, $SNR_\ell(k)$ and M, and maps them to a constrained spectral efficiency $S_\ell(k)$. Here, x represents the set of pertinent variables for the function. The inverse constrained SISO spectral efficiency function $f_{siso}^{-1}(x)$ takes two inputs, S_{avg} and M, and maps them to an SNR value, as follows:

$$SNR_{equiv} = f_{siso}^{-1}(S_{avg}, M) \, . \qquad \text{Eq (11)}$$

The inverse function $f_{siso}^{-1}(S_{avg}, M)$ may be determined once for each supported modulation scheme and stored in a look-up table.

[0049] The highest rate that may be used for data transmission in a practical MIMO-OFDM system with an AWGN channel is then determined based on the equivalent SNR for the equivalent system (block 332). The practical MIMO-OFDM system may support a set of P rates, $R = \{R(m),\ m = 1,\ 2,\ \ldots\ \text{P}\}$, where m is a rate index. Only the P rates in set R are available for use for data transmission. Each rate $R(m)$ in set R may be associated with a specific modulation scheme $M(m)$, a specific code rate or coding scheme $C(m)$, a specific data rate $D(m)$, and a specific required SNR $SNR_{req}(m)$, as follows:

$$R(m) \leftrightarrow [M(m),\ C(m),\ D(m),\ SNR_{req}(m)] \, , \quad \text{for } m = 1 \ \ldots\ \text{P} \, . \qquad \text{Eq (12)}$$

For each rate $R(m)$, the data rate $D(m)$ is determined by the modulation scheme $M(m)$ and the code rate $C(m)$. For example, a rate associated with a modulation scheme of QPSK (with two bits per modulation symbol) and a code rate of 1/2 would have a data rate of 1.0 information bit per modulation symbol. Expression (12) states that data rate $D(m)$ may be transmitted using modulation scheme $M(m)$ and code rate $C(m)$ and further requires an SNR of $SNR_{req}(m)$ or better to achieve a PER of P_e. The required SNR accounts for system losses in the practical system and may be determined by computer simulation, empirical measurements, and so on. The set of supported rates and their required SNRs may be stored in a look-up table. The equivalent SNR SNR_{equiv} may be provided to the look-up table, which then returns the rate $R = R(m_s)$ associated

with the highest data rate supported by SNR_{equiv}. The selected rate R is such that the following conditions are met: (1) modulation scheme M is used for data transmission, or $M(m_s) = M$, (2) the required SNR is less than or equal to the equivalent SNR, or $SNR_{req}(m_s) \leq SNR_{equiv}$, and (3) the maximum data rate is selected, or $D_s = \max_m\{D(m)\}$, subject to the other conditions. The selected rate R includes a back-off factor that accounts for loss due to the selected code rate $C(m_s)$, which may not be able to achieve capacity. This back off occurs in condition (2) above.

[0050] The data rate D_s is indicative of the maximum data rate that can be transmitted on each subband of each spatial channel for a capacity achieving system. An aggregate data rate for all N_T spatial channels may be computed, as follows:

$$D_{total} = D_s \cdot N_T \,. \qquad \text{Eq (13)}$$

The aggregate data rate is given in units of bps/Hz, which is normalized to frequency. The factor of N_F is thus not included in equation (13). The aggregate data rate represents a prediction of the data rate that can be supported by the practical MIMO-OFDM system with the multipath MIMO channel for the desired PER of P_e.

[0051] The rate selection technique described above assumes that the practical MIMO-OFDM system is capable of achieving capacity with modulation scheme M. Several transmission schemes that can achieve capacity are described below. The selected rate R may be an accurate rate for such a system and may be used for data transmission without any modification.

[0052] However, as with any rate prediction scheme, there will inevitably be errors in the rate prediction. Moreover, the practical system may not be able to achieve capacity and/or may have other losses that are unaccounted for by the selected rate R. In this case, to ensure that the desired PER can be achieved, errors in the rate prediction may be estimated and an additional back-off factor may be derived. The rate obtained in block 332 may then be reduced by the additional back-off factor to obtain a final rate for data transmission via the multipath MIMO channel. Alternatively, the average constrained spectral efficiency S_{avg} may be reduced by the additional back-off factor, and the reduced average constrained spectral efficiency may be provided to the look-up table to obtain the rate for data transmission. In any case, the additional back-off factor

reduces the throughput of the system. Thus, it is desirable to keep this back-off factor as small as possible while still achieving the desired PER. An accurate rate prediction scheme, such as the one described herein, may minimize the amount of additional back off to apply and hence maximize system capacity.

[0053] The rate selection described above may be performed continually for each time interval, which may be of any duration (e.g., one OFDM symbol period). It is desirable to use the selected rate for data transmission as soon as possible to minimize the amount of time between the selection of the rate and the use of the rate.

[0054] **FIG. 4A** illustrates the constrained spectral efficiencies for the N_T spatial channels in the MIMO-OFDM system with the multipath MIMO channel. For each spatial channel, a plot 410 of the constrained spectral efficiencies for the N_F subbands may be derived based on the post-detection SNRs, the modulation scheme M, and the constrained SISO spectral efficiency function $f_{siso}(SNR_\ell(k),\ M)$, as shown in equations (8) and (9). Plots 410a through 410t for the N_T spatial channels may be different because of different fading for these spatial channels, as shown in FIG. 4A.

[0055] **FIG. 4B** illustrates the constrained spectral efficiency of the equivalent system with the AWGN channel. A plot 420 is formed by concatenation of plots 410a through 410t for the N_T spatial channels in FIG. 4A. A plot 422 shows the constrained spectral efficiency for the equivalent system, which is the average of the constrained spectral efficiencies for plots 410a through 410t.

[0056] The rate selection described above includes a back-off factor for code loss but otherwise assumes that the MIMO-OFDM system can achieve capacity. Two exemplary transmission schemes capable of achieving capacity are described below.

[0057] In a first transmission scheme, the transmitter transmits data on "eigenmodes" of the MIMO channel. The eigenmodes may be viewed as orthogonal spatial channels obtained by decomposing the MIMO channel. The channel response matrix $\underline{\mathbf{H}}(k)$ for each subband may be decomposed using eigenvalue decomposition, as follows:

$$\underline{\mathbf{R}}(k) = \underline{\mathbf{H}}^H(k) \cdot \underline{\mathbf{H}}(k) = \underline{\mathbf{E}}(k) \cdot \underline{\Lambda}(k) \cdot \underline{\mathbf{E}}^H(k)\ , \quad \text{for } k = 1\ \dots\ N_F, \qquad \text{Eq (14)}$$

where $\underline{\mathbf{R}}(k)$ is an $N_T \times N_T$ correlation matrix of $\underline{\mathbf{H}}(k)$;

$\underline{\mathbf{E}}(k)$ is an $N_T \times N_T$ unitary matrix whose columns are eigenvectors of $\underline{\mathbf{R}}(k)$; and

$\underline{\Lambda}(k)$ is an $N_T \times N_T$ diagonal matrix of eigenvalues of $\underline{\mathbf{R}}(k)$.

A unitary matrix $\underline{\mathbf{U}}$ is characterized by the property $\underline{\mathbf{U}}^H \cdot \underline{\mathbf{U}} = \underline{\mathbf{I}}$. The columns of a unitary matrix are orthogonal to one another.

[0058] The transmitter performs spatial processing as follows:

$$\underline{\mathbf{x}}(k) = \underline{\mathbf{E}}(k) \cdot \underline{\mathbf{s}}(k) \ , \quad \text{for } k = 1 \ \ldots \ \mathrm{N_F}, \qquad \text{Eq (15)}$$

where $\underline{\mathbf{s}}(k)$ is an $\mathrm{N_T} \times 1$ vector with $\mathrm{N_T}$ data symbols to be sent on the $\mathrm{N_T}$ eigenmodes of subband k; and

$\underline{\mathbf{x}}(k)$ is an $\mathrm{N_T} \times 1$ vector with $\mathrm{N_T}$ transmit symbols to be transmitted from the $\mathrm{N_T}$ transmit antennas on subband k.

[0059] The received symbols at the receiver may be expressed as:

$$\underline{\mathbf{r}}_{em}(k) = \underline{\mathbf{H}}(k) \cdot \underline{\mathbf{x}}(k) + \underline{\mathbf{n}}(k) \ , \quad \text{for } k = 1 \ \ldots \ \mathrm{N_F}, \qquad \text{Eq (16)}$$

where $\underline{\mathbf{r}}_{em}(k)$ is an $\mathrm{N_R} \times 1$ vector with $\mathrm{N_R}$ received symbols obtained via the $\mathrm{N_R}$ receive antennas on subband k; and

$\underline{\mathbf{n}}$ is an $\mathrm{N_R} \times 1$ vector of noise and interference for subband k.

The noise vector $\underline{\mathbf{n}}(k)$ is assumed to have zero mean and a covariance matrix of $\underline{\mathbf{\Lambda}}_n(k) = N_0 \cdot \underline{\mathbf{I}}$, where N_0 is the noise variance.

[0060] The receiver performs receiver spatial processing/detection, as follows:

$$\hat{\underline{\mathbf{s}}}_{em}(k) = \underline{\mathbf{\Lambda}}^{-1}(k) \cdot \underline{\mathbf{E}}^H(k) \cdot \underline{\mathbf{H}}^H(k) \cdot \underline{\mathbf{r}}_{em}(k) = \underline{\mathbf{s}}(k) + \underline{\mathbf{n}}_{em}(k) \ , \quad k = 1 \ \ldots \ \mathrm{N_F}, \qquad \text{Eq (17)}$$

where $\hat{\underline{\mathbf{s}}}_{em}(k)$ is an $\mathrm{N_T} \times 1$ vector with $\mathrm{N_T}$ detected symbols for subband k, which are estimates of the $\mathrm{N_T}$ data symbols in $\underline{\mathbf{s}}(k)$; and

$\underline{\mathbf{n}}_{em}(k) = \underline{\mathbf{\Lambda}}^{-1}(k) \cdot \underline{\mathbf{E}}^H(k) \cdot \underline{\mathbf{H}}^H(k) \cdot \underline{\mathbf{n}}(k)$ is the post-detection interference and noise after the spatial processing at the receiver.

Each eigenmode is an effective channel between an element of the data symbol vector $\underline{\mathbf{s}}(k)$ and a corresponding element of the detected symbol vector $\hat{\underline{\mathbf{s}}}_{em}(k)$.

[0061] The SNR for each subband of each eigenmode may be expressed as:

$$SNR_{em,\ell}(k) = \frac{P_\ell(k) \cdot \lambda_\ell(k)}{N_0} \ , \quad \text{for } k = 1 \ \ldots \ \mathrm{N_F} \text{ and } \ell = 1 \ \ldots \ \mathrm{N_T}, \qquad \text{Eq (18)}$$

where $P_\ell(k)$ is the transmit power used for eigenmode ℓ of subband k;

$\lambda_\ell(k)$ is the eigenvalue for eigenmode ℓ of subband k, which is the ℓ-th diagonal element of $\underline{\Lambda}(k)$; and

$SNR_{em,\ell}(k)$ is the post-detection SNR for eigenmode ℓ of subband k.

[0062] In a second transmission scheme, the transmitter encodes and modulates data to obtain data symbols, demultiplexes the data symbols into N_T data symbol streams, and transmits the N_T data symbol streams simultaneously from the N_T transmit antennas. The received symbols at the receiver may be expressed as:

$$\underline{r}_{am}(k) = \underline{\mathbf{H}}(k) \cdot \underline{s}(k) + \underline{n}(k) \ , \quad \text{for } k = 1 \ldots N_F . \qquad \text{Eq (19)}$$

[0063] The receiver performs receiver spatial processing/detection on the N_R received symbols for each subband to recover the N_T data symbols transmitted on that subband. The receiver spatial processing may be performed with a minimum mean square error (MMSE) detector, a maximal ratio combining (MRC) detector, a linear zero-forcing (ZF) detector, an MMSE linear equalizer (MMSE-LE), a decision feedback equalizer (DFE), or some other detector/equalizer.

[0064] The receiver may also process the N_R received symbol streams using a successive interference cancellation (SIC) technique to recover the N_T data symbol streams. The SIC technique may be used when the transmitter independently processes the N_T data symbol streams so that the receiver can individually recover each data symbol stream. The receiver recovers the N_T data symbol streams in N_T successive stages, one data symbol stream in each stage.

[0065] For the first stage, the receiver initially performs receiver spatial processing/detection on the N_R received symbol streams (e.g., using an MMSE, MRC, or zero-forcing detector) and obtains one detected symbol stream. The receiver further demodulates, deinterleaves, and decodes the detected symbol stream to obtain a decoded data stream. The receiver then estimates the interference this decoded data stream causes to the other $N_T - 1$ data symbol streams not yet recovered, cancels the estimated interference from the N_R received symbol streams, and obtains N_R modified symbol streams for the next stage. The receiver then repeats the same processing on the N_R modified symbol streams to recover another data symbol stream. For simplicity, the following description assumes that the N_T data symbol streams are recovered in

sequential order, i.e., data symbol stream $\{s_\ell(k)\}$ sent from transmit antenna ℓ is recovered in the ℓ-th stage, for $\ell = 1 \ldots N_T$.

[0066] For a SIC with MMSE receiver, an MMSE detector is derived for each subband of stage ℓ, for $\ell = 1 \ldots N_T$, as follows:

$$\underline{\mathbf{W}}_{mmse,\ell}(k) = (\underline{\mathbf{H}}_\ell(k) \cdot \underline{\mathbf{H}}_\ell^H(k) + N_0 \cdot \underline{\mathbf{I}})^{-1} \cdot \underline{\mathbf{H}}_\ell(k) \ , \quad \text{for } k = 1 \ldots N_F, \qquad \text{Eq (20)}$$

where $\underline{\mathbf{W}}_{mmse,\ell}(k)$ is an $N_R \times (N_T - \ell + 1)$ matrix for the MMSE detector for subband k in stage ℓ; and

$\underline{\mathbf{H}}_\ell(k)$ is an $N_R \times (N_T - \ell + 1)$ reduced channel response matrix for subband k in stage ℓ.

The reduced channel response matrix $\underline{\mathbf{H}}_\ell(k)$ is obtained by removing $\ell - 1$ columns in the original matrix $\underline{\mathbf{H}}(k)$ corresponding to the $\ell - 1$ data symbol streams already recovered in the $\ell - 1$ prior stages.

[0067] The receiver performs detection for each subband in stage ℓ, as follows:

$$\hat{s}_{mmse,\ell}(k) = \underline{\mathbf{w}}_{mmse,\ell}^H(k) \cdot \underline{\mathbf{r}}_\ell(k) = s_\ell(k) + \underline{\mathbf{w}}_{mmse,\ell}^H(k) \cdot \underline{\mathbf{n}}_\ell(k) \ , \qquad \text{Eq (21)}$$

where $\underline{\mathbf{w}}_{mmse,\ell}(k)$ is a column of $\underline{\mathbf{W}}_{mmse,\ell}(k)$ corresponding to transmit antenna ℓ;

$\hat{s}_{mmse,\ell}(k)$ is the MMSE detected symbol for subband k in stage ℓ; and

$\underline{\mathbf{w}}_{mmse,\ell}^H(k) \cdot \underline{\mathbf{n}}_\ell(k)$ is the post-detection noise for the detected symbol $\hat{s}_{mmse,\ell}(k)$.

[0068] The SNR for each subband of each transmit antenna may be expressed as:

$$SNR_{mmse,\ell}(k) = \frac{P_\ell(k)}{N_0 \cdot \| \underline{\mathbf{w}}_{mmse,\ell}(k) \|^2} \ , \qquad \text{Eq (22)}$$

where $N_0 \cdot \| \underline{\mathbf{w}}_{mmse,\ell}(k) \|^2$ is the variance of the post-detection noise; and

$SNR_{mmse,\ell}(k)$ is the post-detection SNR for subband k of transmit antenna ℓ.

The post-detection SNRs for later stages improve because the norm of $\underline{\mathbf{w}}_{mmse,\ell}(k)$ in equation (22) decreases with each stage.

[0069] The SIC technique is described in further detail in commonly assigned U.S. Patent Application Serial No. 09/993,087, entitled "Multiple-Access Multiple-Input Multiple-Output (MIMO) Communication System," filed November 6, 2001.

[0070] For the second transmission scheme, the receiver can also recover the N_T data symbol streams using an iterative detection and decoding (IDD) scheme. For the IDD scheme, whenever a block of received symbols for a data packet is obtained, the receiver iteratively performs detection and decoding multiple (N_{dec}) times on the received symbols in the block to obtain a decoded packet. A detector performs detection on the received symbol block and provides a detected symbol block. A decoder performs decoding on the detected symbol block and provides decoder *a priori* information, which is used by the detector in a subsequent iteration. The decoded packet is generated based on decoder output for the last iteration.

[0071] It can be shown that the first transmission scheme and the second transmission scheme with either the SIC with MMSE receiver or the IDD receiver are optimal and can achieve capacity or near capacity for the MIMO-OFDM system. The second transmission scheme with a maximum likelihood detector for the received symbols can also provide optimal or near optimal performance. Other capacity achieving transmission schemes may also be used for the MIMO-OFDM system. One such capacity achieving transmission scheme is an autocoding scheme described by T.L. Marzetta *et al.* in a paper entitled "Structured Unitary Space-Time Autocoding Constellations," IEEE Transaction on Information Theory, Vol. 48, No. 4, April 2002.

[0072] **FIG. 5** shows a block diagram of transmitter 110. Within TX data processor 120, an encoder 520 receives and encodes a data stream $\{d\}$ in accordance with coding scheme C for the selected rate R and provides code bits. The encoding increases the reliability of the data transmission. The coding scheme may include a convolutional code, a Turbo code, a block code, a CRC code, or a combination thereof. A channel interleaver 522 interleaves (i.e., reorders) the code bits from encoder 520 based on an interleaving scheme. The interleaving provides time and/or frequency diversity for the code bits. A symbol mapping unit 524 modulates (i.e., symbol maps) the interleaved data from channel interleaver 522 in accordance with modulation scheme M for the selected rate R and provides data symbols. The modulation may be achieved by (1) grouping sets of B interleaved bits to form B-bit binary values, where $B \geq 1$, and (2) mapping each B-bit binary value to a specific signal point in a signal constellation for

the modulation scheme. Symbol mapping unit 524 provides a stream of data symbols $\{s\}$.

[0073] Transmitter 110 encodes and modulates each data packet separately based on the rate R selected for the packet to obtain a corresponding data symbol block. Transmitter 110 may transmit one data symbol block at a time on all subbands of all spatial channels available for data transmission. Each data symbol block may be transmitted in one or multiple OFDM symbol periods. Transmitter 110 may also transmit multiple data symbol blocks simultaneously on the available subbands and spatial channels. If one rate is selected for each time interval, as described above, then all data symbol block(s) transmitted in the same time interval use the same selected rate.

[0074] For the embodiment shown in FIG. 5, TX spatial processor 130 implements the second transmission scheme described above. Within TX spatial processor 130, a multiplexer/demultiplexer (Mux/Demux) 530 receives and demultiplexes the data symbol stream $\{s\}$ into N_T streams for the N_T transmit antennas. Mux/demux 530 also multiplexes in pilot symbols (e.g., in a time division multiplex (TDM) manner) and provides N_T transmit symbol streams, $\{x_1\}$ through $\{x_{N_T}\}$, for the N_T transmit antennas. Each transmit symbol may be a data symbol, a pilot symbol, or a signal value of zero for a subband not used for data or pilot transmission.

[0075] Transmitter unit 132 includes N_T OFDM modulators 532a through 532t and N_T TX RF units 534a through 534t for the N_T transmit antennas. Each OFDM modulator 532 performs OFDM modulation on a respective transmit symbol stream by (1) grouping and transforming each set of N_F transmit symbols for the N_F subbands to the time domain using an N_F-point IFFT to obtain a corresponding transformed symbol that contains N_F chips and (2) repeating a portion (or N_{cp} chips) of each transformed symbol to obtain a corresponding OFDM symbol that contains $N_F + N_{cp}$ chips. The repeated portion is referred to as a cyclic prefix, which ensures that the OFDM symbol retains its orthogonal properties in the presence of delay spread in a multipath channel. Each OFDM modulator 532 provides a stream of OFDM symbols, which is further conditioned (e.g., converted to analog, frequency upconverted, filtered, and amplified) by an associated TX RF unit 534 to generate a modulated signal. The N_T modulated signals from TX RF units 534a through 534t are transmitted from N_T antennas 540a through 540t, respectively.

[0076] FIG. 6 shows a block diagram of receiver 150. N_R receive antennas 652a through 652r receive the modulated signals transmitted by transmitter 110 and provide N_R received signals to receiver unit 154. Receiver unit 154 includes N_R RX RF units 654a through 654r and N_R OFDM demodulators 656a through 656r for the N_R receive antennas. Each RX RF unit 654 conditions and digitizes a respective received signal and provides a stream of samples. Each OFDM demodulator 656 performs OFDM demodulation on a respective sample stream by (1) removing the cyclic prefix in each received OFDM symbol to obtain a received transformed symbol and (2) transforming each received transformed symbol to the frequency domain with an N_F-point FFT to obtain N_F received symbols for the N_F subbands. Each OFDM demodulator 656 provides received data symbols to RX spatial processor 160 and received pilot symbols to channel estimator 172.

[0077] FIG. 6 also shows an RX spatial processor 160a and an RX data processor 170a, which are one embodiment of RX spatial processor 160 and RX data processor 170, respectively, at receiver 150. Within RX spatial processor 160a, a detector 660 performs spatial processing/detection on the N_R received symbol streams to obtain N_T detected symbol streams. Each detected symbol is an estimate of a data symbol transmitted by the transmitter. Detector 660 may implement an MMSE, MRC, or zero-forcing detector. The detection is performed for each subband based on a matched filter matrix (or detector response) $\underline{\mathbf{W}}(k)$ for that subband, which is derived based on an estimate of the channel response matrix $\underline{\mathbf{H}}(k)$ for the subband. For example, the matched filter matrix for the MMSE detector may be derived as: $\underline{\mathbf{W}}_{mmse}(k) = (\underline{\mathbf{H}}(k) \cdot \underline{\mathbf{H}}^H(k) + N_0 \cdot \underline{\mathbf{I}})^{-1} \cdot \underline{\mathbf{H}}(k)$. A multiplexer 662 multiplexes the detected symbols and provides a detected symbol stream $\{\hat{s}\}$ to RX data processor 170a.

[0078] Within RX data processor 170a, a symbol demapping unit 670 demodulates the detected symbols in accordance with the modulation scheme M for the selected rate R and provides demodulated data. A channel deinterleaver 672 deinterleaves the demodulated data in a manner complementary to the interleaving performed at the transmitter and provides deinterleaved data. A decoder 674 decodes the deinterleaved data in a manner complementary to the encoding performed at the transmitter and provides a decoded data stream $\{\hat{d}\}$. For example, decoder 674 may implement a Turbo decoder or a Viterbi decoder if Turbo or convolutional coding, respectively, is

performed at the transmitter. Decoder 674 also provides the status of each decoded packet, which indicates whether the packet is decoded correctly or in error.

[0079] FIG. 7 shows an RX spatial processor 160b and an RX data processor 170b, which implement the IDD scheme and are another embodiment of RX spatial processor 160 and RX data processor 170, respectively, at receiver 150. A detector 760 and a decoder 780 perform iterative detection and decoding on the received symbols for each data packet to obtain a decoded packet. The IDD scheme exploits the error correction capabilities of the channel code to provide improved performance. This is achieved by iteratively passing *a priori* information between detector 760 and decoder 780 for N_{dec} iterations, where $N_{dec} > 1$. The *a priori* information indicates the likelihood of each transmitted data bit being zero or one.

[0080] Within RX spatial processor 160b, a buffer 758 receives and stores N_R received symbol sequences from the N_R receive antennas for each data packet. The iterative detection and decoding process is performed on each block of received symbols for a data packet. Detector 760 performs spatial processing on the N_R received symbol sequences for each block and provides N_T detected symbol sequences for the block. Detector 760 may implement an MMSE, MRC, or zero-forcing detector. A multiplexer 762 multiplexes the detected symbols in the N_T sequences and provides a detected symbol block.

[0081] Within RX data processor 170b, a log-likelihood ratio (LLR) computation unit 770 receives the detected symbols from RX spatial processor 160b and computes the LLRs for the B code bits of each detected symbol. These LLRs represent *a priori* information provided by detector 760 to decoder 780. A channel deinterleaver 772 deinterleaves each block of LLRs from LLR computation unit 770 and provides deinterleaved LLRs $\{x^n\}$ for the block. Decoder 780 decodes the deinterleaved LLRs and provides decoder LLRs $\{x^{n+1}\}$, which represent *a priori* information provided by decoder 780 to detector 760. The decoder LLRs are interleaved by a channel interleaver 782 and provided to detector 760.

[0082] The detection and decoding process is then repeated for another iteration. Detector 760 derives new detected symbols based on the received symbols and the decoder LLRs. The new detected symbols are again decoded by decoder 780. The detection and decoding process is iterated N_{dec} times. During the iterative detection and decoding process, the reliability of the detected symbols improves with each

detection/decoding iteration. After all N_{dec} detection/decoding iterations have been completed, decoder 780 computes the final data bit LLRs and slices these LLRs to obtain the decoded packet.

[0083] The IDD scheme is described in further detail in commonly assigned U.S. Patent Application Serial No. 60/506,466, entitled "Hierarchical Coding With Multiple Antennas in a Wireless Communication System," filed September 25, 2003.

[0084] The rate selection and data transmission techniques described herein may be implemented by various means. For example, these techniques may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the processing units used to perform rate selection and data transmission may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[0085] For a software implementation, the rate selection and data transmission techniques may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memory unit 182 or 142 in FIG. 1) and executed by a processor (e.g., controller 180 or 140). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as is known in the art.

[0086] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[0087] **WHAT IS CLAIMED IS:**

CLAIMS

1. A method of selecting a rate for data transmission in a multi-carrier multiple-input multiple-output (MIMO) communication system, comprising:

determining an average constrained spectral efficiency for a plurality of subbands of a plurality of spatial channels used for data transmission, the plurality of spatial channels being formed by a MIMO channel in the system;

determining an equivalent signal-to-noise-and-interference ratio (SNR) needed by an equivalent system with an additive white Gaussian noise (AWGN) channel to support the average constrained spectral efficiency; and

selecting the rate for data transmission in the multi-carrier MIMO system based on the equivalent SNR.

2. The method of claim 1, wherein the average constrained spectral efficiency, the equivalent SNR, and the rate are all determined based on a specific modulation scheme.

3. The method of claim 1, wherein the plurality of subbands are obtained with orthogonal frequency division multiplexing (OFDM).

4. The method of claim 1, wherein the plurality of spatial channels correspond to a plurality of single-input single-output (SISO) channels that make up the MIMO channel.

5. The method of claim 1, further comprising:

determining a post-detection SNR for each subband of each spatial channel used for data transmission; and

determining a constrained spectral efficiency for each subband of each spatial channel based on the post-detection SNR for the subband of the spatial channel, and

wherein the average constrained spectral efficiency is determined based on constrained spectral efficiencies for the plurality of subbands of the plurality of spatial channels.

6. The method of claim 5, wherein the post-detection SNR for each subband of each spatial channel is determined based on a transmission scheme capable of achieving capacity of the MIMO channel.

7. The method of claim 5, wherein the post-detection SNR for each subband of each spatial channel is determined based on successive interference cancellation (SIC) processing with a minimum mean square error (MMSE) detector at a receiver.

8. The method of claim 5, wherein the constrained spectral efficiency for each subband of each spatial channel is further determined based on a constrained spectral efficiency function having an SNR and a modulation scheme as inputs and providing a constrained spectral efficiency as output.

9. The method of claim 1, further comprising:

determining a constrained spectral efficiency for each subband of the MIMO channel based on a constrained spectral efficiency function having a MIMO channel response and a modulation scheme as inputs and providing a constrained spectral efficiency as output, and

wherein the average constrained spectral efficiency for the plurality of subbands of the plurality of spatial channels used for data transmission is determined based on constrained spectral efficiencies for the plurality of subbands of the MIMO channel.

10. The method of claim 1, wherein the equivalent SNR is determined based on an inverse constrained spectral efficiency function having a spectral efficiency and a modulation scheme as inputs and providing an SNR as output.

11. The method of claim 1, wherein the rate for data transmission is selected based on a set of rates supported by the multi-carrier MIMO system and required SNRs for the supported rates.

12. The method of claim 11, wherein the selected rate is a highest rate among the supported rates having a required SNR less than or equal to the equivalent SNR.

13. The method of claim 11, wherein the required SNRs for the supported rates include losses observed by the multi-carrier MIMO system.

14. The method of claim 1, further comprising:

determining a back-off factor to account for error in rate prediction and system losses; and

reducing the rate for data transmission based on the back-off factor.

15. The method of claim 1, further comprising:

receiving a data transmission at the selected rate, wherein the received data transmission includes at least one block of data symbols for at least one data packet, and wherein the data symbols in each block are transmitted simultaneously on the plurality of subbands of the plurality of spatial channels used for data transmission.

16. The method of claim 1, further comprising:

receiving a data transmission at the selected rate; and

performing iterative detection and decoding (IDD) to recover data in the received data transmission.

17. An apparatus in a multi-carrier multiple-input multiple-output (MIMO) communication system, comprising:

a channel estimator operative to obtain channel estimates for a MIMO channel in the system; and

a controller operative to

determine an average constrained spectral efficiency for a plurality of subbands of a plurality of spatial channels used for data transmission based on the channel estimates, wherein the plurality of spatial channels are formed by the MIMO channel,

determine an equivalent signal-to-noise-and-interference ratio (SNR) needed by an equivalent system with an additive white Gaussian noise (AWGN) channel to support the average constrained spectral efficiency, and

select a rate for data transmission in the multi-carrier MIMO system based on the equivalent SNR.

18. The apparatus of claim 17, wherein the controller is further operative to

determine a post-detection SNR for each subband of each spatial channel used for data transmission based on the channel estimates, and

determine a constrained spectral efficiency for each subband of each spatial channel based on the post-detection SNR for the subband of the spatial channel, and wherein the average constrained spectral efficiency is determined based on constrained spectral efficiencies for the plurality of subbands of the plurality of spatial channels.

19. The apparatus of claim 18, wherein the post-detection SNR for each subband of each spatial channel is further determined based on a transmission scheme capable of achieving capacity of the MIMO channel.

20. The apparatus of claim 17, wherein a set of rates is supported by the multi-carrier MIMO system and each supported rate is associated with a respective required SNR, and wherein the controller is further operative to select a highest rate among the supported rates having a required SNR less than or equal to the equivalent SNR.

21. The apparatus of claim 17, wherein the controller is further operative to determine a back-off factor to account for error in rate prediction and system losses and to reduce the rate for data transmission based on the back-off factor.

22. The apparatus of claim 17, further comprising:

a receive spatial processor operative to perform detection on received symbols for a data transmission at the selected rate and provide detected symbols; and

a receive data processor operative to process the detected symbols to obtain decoded data.

23. The apparatus of claim 22, wherein the receive spatial processor and the receive data processor are operative to perform iterative detection and decoding (IDD) to obtain the decoded data from the received symbols.

24. An apparatus in a multi-carrier multiple-input multiple-output (MIMO) communication system, comprising:

means for determining an average constrained spectral efficiency for a plurality of subbands of a plurality of spatial channels used for data transmission, the plurality of spatial channels being formed by a MIMO channel in the system;

means for determining an equivalent signal-to-noise-and-interference ratio (SNR) needed by an equivalent system with an additive white Gaussian noise (AWGN) channel to support the average constrained spectral efficiency; and

means for selecting a rate for data transmission in the multi-carrier MIMO system based on the equivalent SNR.

25. The apparatus of claim 24, further comprising:

means for determining a post-detection SNR for each subband of each spatial channel used for data transmission; and

means for determining a constrained spectral efficiency for each subband of each spatial channel based on the post-detection SNR for the subband of the spatial channel, and wherein the average constrained spectral efficiency is determined based on constrained spectral efficiencies for the plurality of subbands of the plurality of spatial channels.

26. The apparatus of claim 24, further comprising:

means for determining a back-off factor to account for error in rate prediction and system losses; and

means for reducing the rate for data transmission based on the back-off factor.

27. The apparatus of claim 24, further comprising:

means for receiving a data transmission at the selected rate; and

means for performing iterative detection and decoding (IDD) to recover data in the received data transmission.

28. A processor readable media for storing instructions operable in an apparatus to:

determine an average constrained spectral efficiency for a plurality of subbands of a plurality of spatial channels used for data transmission in a multi-carrier multiple-

input multiple-output (MIMO) communication system, the plurality of spatial channels being formed by a MIMO channel in the system;

determine an equivalent signal-to-noise-and-interference ratio (SNR) needed by an equivalent system with an additive white Gaussian noise (AWGN) channel to support the average constrained spectral efficiency; and

select a rate for data transmission in the multi-carrier MIMO system based on the equivalent SNR.

29. The processor readable media of claim 28 and further for storing instructions operable to

determine a post-detection SNR for each subband of each spatial channel used for data transmission; and

determine a constrained spectral efficiency for each subband of each spatial channel based on the post-detection SNR for the subband of the spatial channel, and wherein the average constrained spectral efficiency is determined based on constrained spectral efficiencies for the plurality of subbands of the plurality of spatial channels.

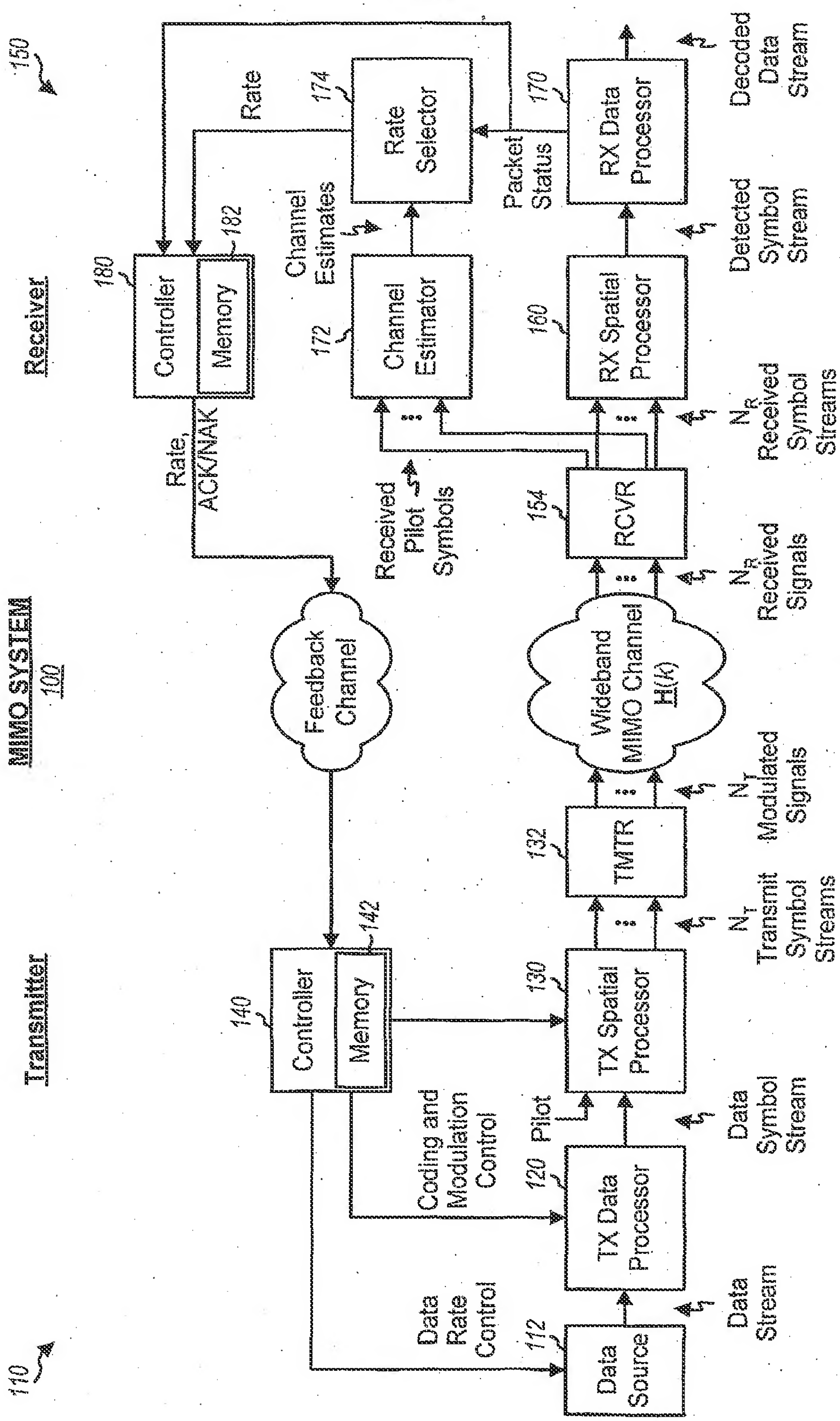

MIMO SYSTEM
100
Transmitter
110
Receiver
150
Data Source
112
Data Rate Control
Data Stream
TX Data Processor
120
Coding and Modulation Control
Pilot
Data Symbol Stream
TX Spatial Processor
130
N_T Transmit Symbol Streams
TMTR
132
N_T Modulated Signals
Controller
Memory
140
142
Feedback Channel
Wideband MIMO Channel $\underline{H}(k)$
N_R Received Signals
RCVR
154
N_R Received Symbol Streams
Received Pilot Symbols
Rate, ACK/NAK
Controller
Memory
180
182
Channel Estimator
172
Channel Estimates
RX Spatial Processor
160
Detected Symbol Stream
Rate Selector
174
Rate
Packet Status
RX Data Processor
170
Decoded Data Stream

FIG. 1

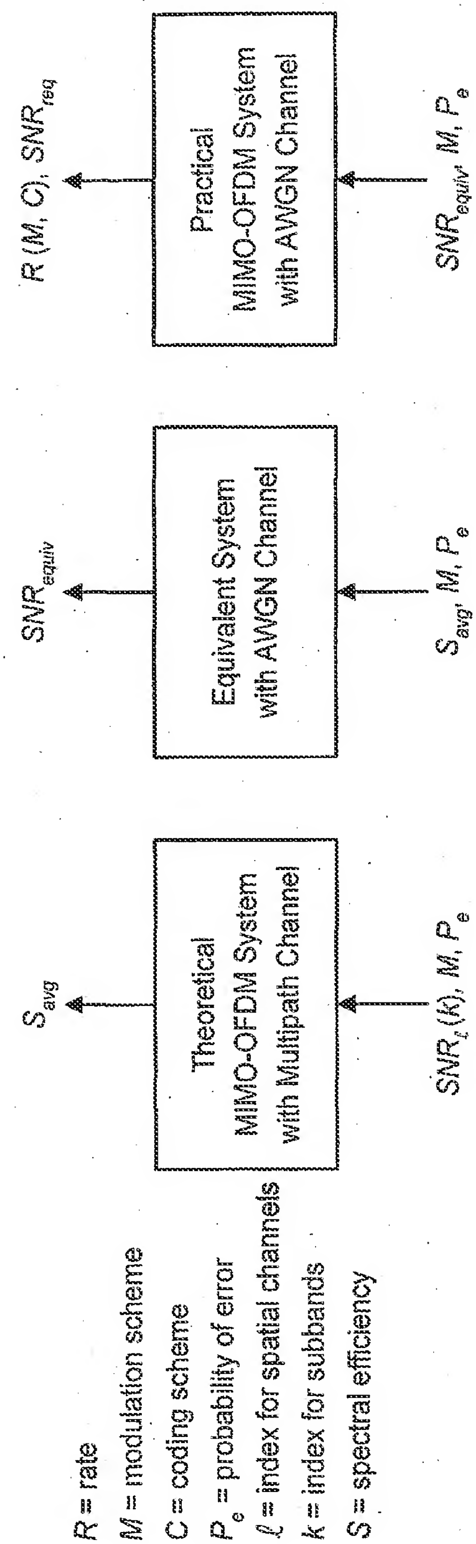

R (M, C), SNR_req
Practical MIMO-OFDM System with AWGN Channel
SNR_equiv, M, P_e
SNR_equiv
Equivalent System with AWGN Channel
S_avg, M, P_e
S_avg
Theoretical MIMO-OFDM System with Multipath Channel
SNR_ℓ (k), M, P_e
R = rate
M = modulation scheme
C = coding scheme
P_e = probability of error
ℓ = index for spatial channels
k = index for subbands
S = spectral efficiency

FIG. 2

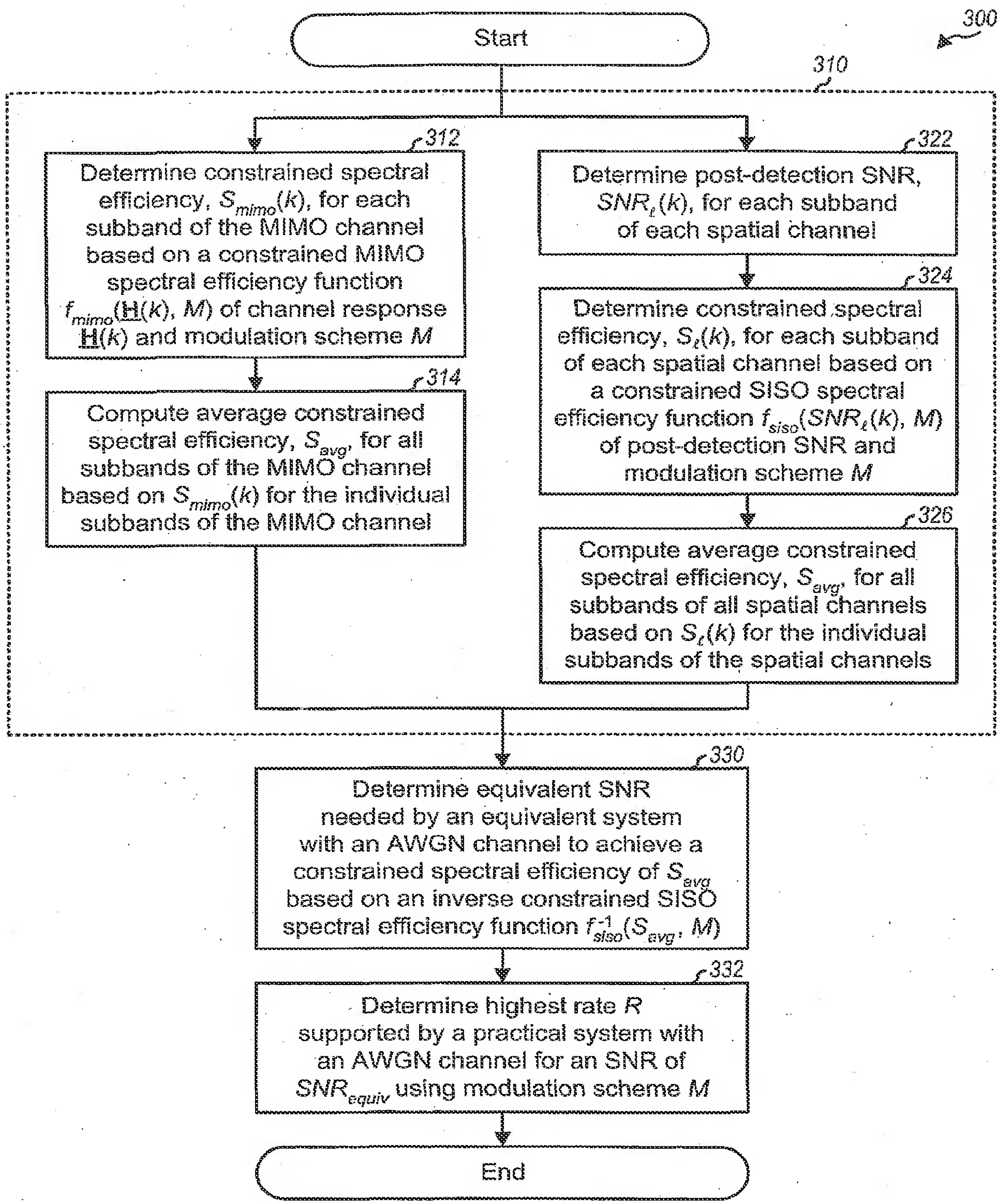
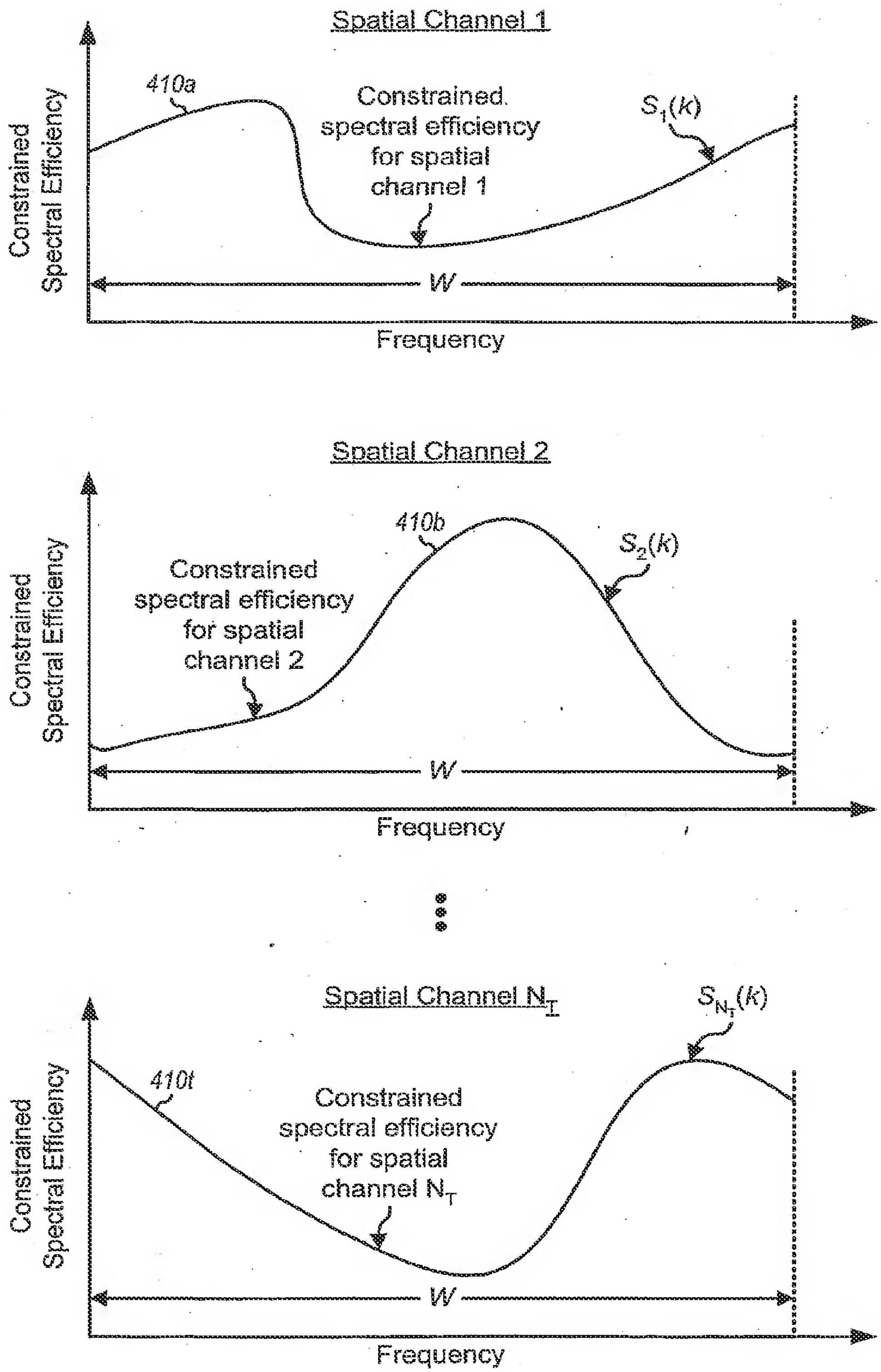

300
Start
310
312
Determine constrained spectral efficiency, $S_{mimo}(k)$, for each subband of the MIMO channel based on a constrained MIMO spectral efficiency function $f_{mimo}(\underline{\mathbf{H}}(k), M)$ of channel response $\underline{\mathbf{H}}(k)$ and modulation scheme M
314
Compute average constrained spectral efficiency, S_{avg}, for all subbands of the MIMO channel based on $S_{mimo}(k)$ for the individual subbands of the MIMO channel
322
Determine post-detection SNR, $SNR_{\ell}(k)$, for each subband of each spatial channel
324
Determine constrained spectral efficiency, $S_{\ell}(k)$, for each subband of each spatial channel based on a constrained SISO spectral efficiency function $f_{siso}(SNR_{\ell}(k), M)$ of post-detection SNR and modulation scheme M
326
Compute average constrained spectral efficiency, S_{avg}, for all subbands of all spatial channels based on $S_{\ell}(k)$ for the individual subbands of the spatial channels
330
Determine equivalent SNR needed by an equivalent system with an AWGN channel to achieve a constrained spectral efficiency of S_{avg} based on an inverse constrained SISO spectral efficiency function $f_{siso}^{-1}(S_{avg}, M)$
332
Determine highest rate R supported by a practical system with an AWGN channel for an SNR of SNR_{equiv} using modulation scheme M
End

FIG. 3



Spatial Channel 1
Constrained Spectral Efficiency
410a
Constrained spectral efficiency for spatial channel 1
$S_1(k)$
W
Frequency
Spatial Channel 2
Constrained Spectral Efficiency
410b
Constrained spectral efficiency for spatial channel 2
$S_2(k)$
W
Frequency
Spatial Channel N_T
Constrained Spectral Efficiency
410t
Constrained spectral efficiency for spatial channel N_T
$S_{N_T}(k)$
W
Frequency

FIG. 4A

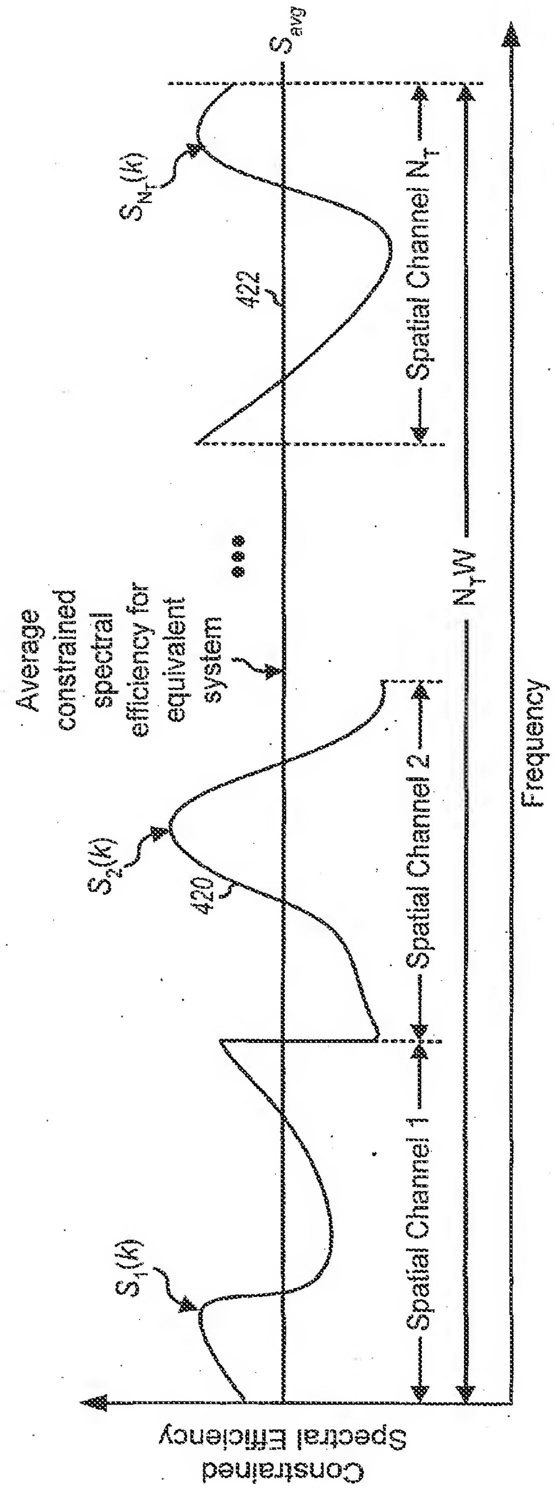

Constrained Spectral Efficiency
Frequency
$S_1(k)$
$S_2(k)$
$S_{N_T}(k)$
420
422
S_{avg}
Average constrained spectral efficiency for equivalent system
Spatial Channel 1
Spatial Channel 2
Spatial Channel N_T
$N_T W$

FIG. 4B

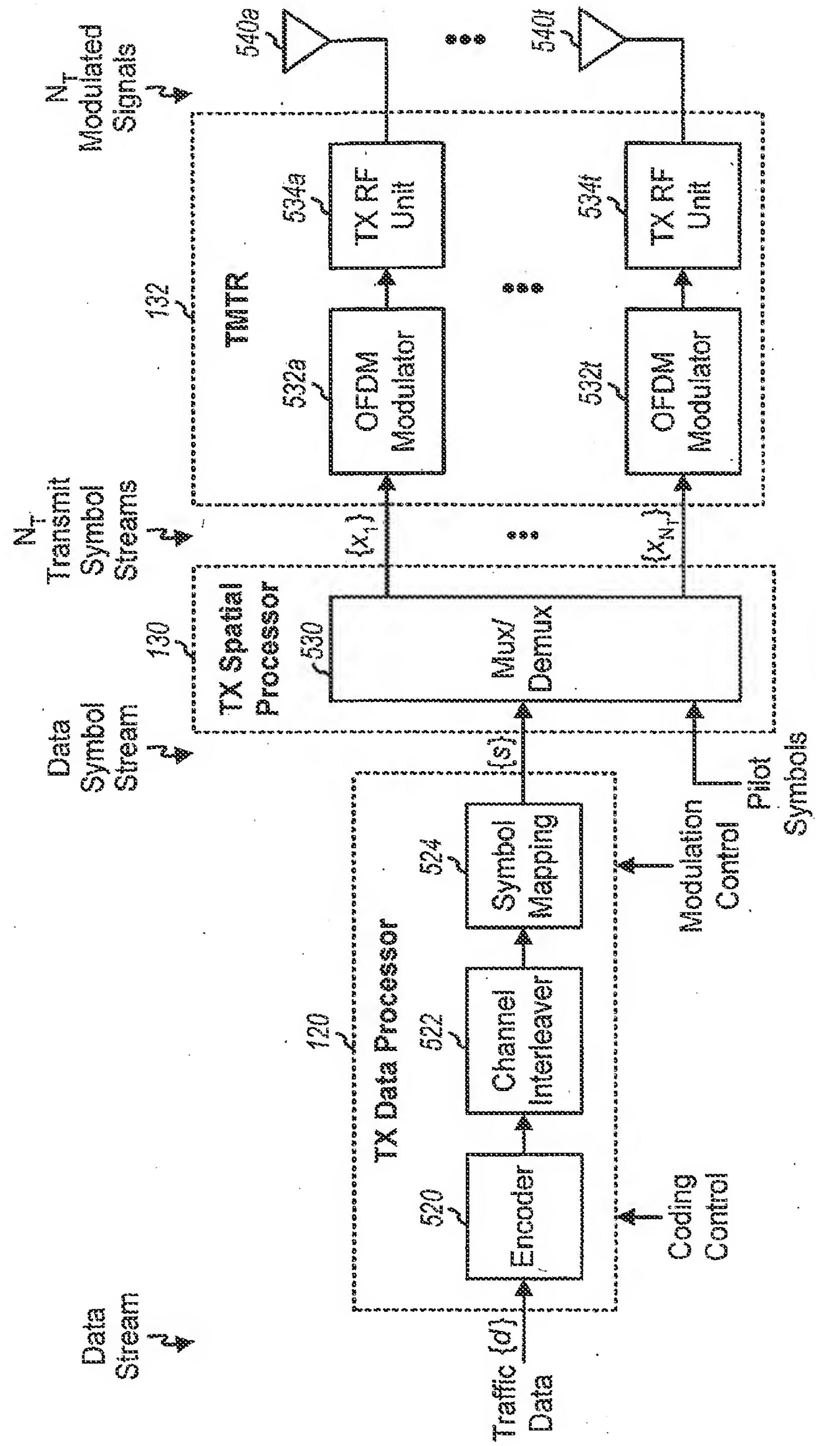

Data Stream
Data Symbol Stream
N_T Transmit Symbol Streams
N_T Modulated Signals
Traffic {d} Data
120
TX Data Processor
520
Encoder
522
Channel Interleaver
524
Symbol Mapping
{s}
Coding Control
Modulation Control
Pilot Symbols
130
TX Spatial Processor
530
Mux/ Demux
$\{x_1\}$
$\{x_{N_T}\}$
132
TMTR
532a
OFDM Modulator
534a
TX RF Unit
540a
532t
OFDM Modulator
534t
TX RF Unit
540t

FIG. 5

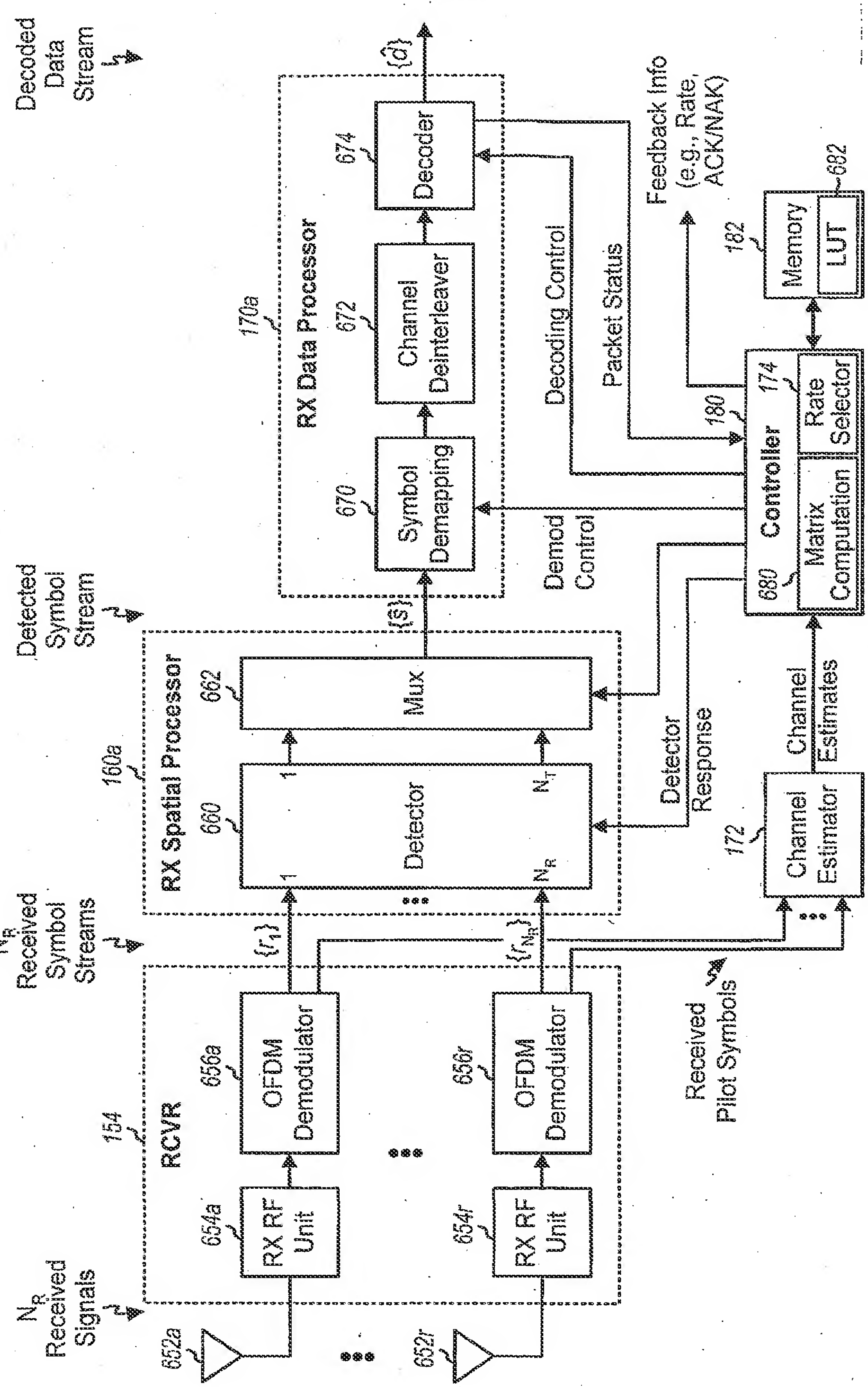
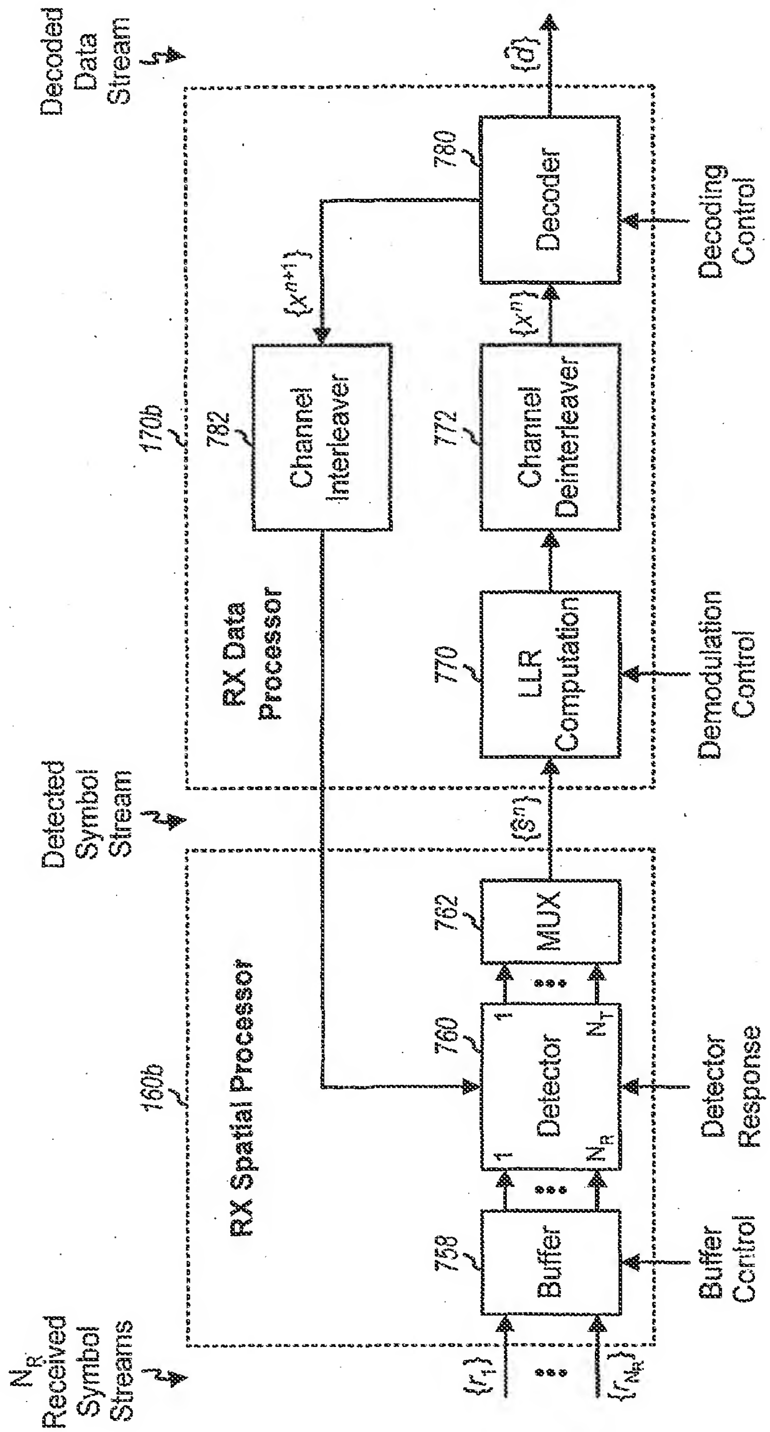

Decoded Data Stream
Detected Symbol Stream
N_R Received Symbol Streams
N_R Received Signals
652a
652r
154
RCVR
654a
RX RF Unit
654r
RX RF Unit
656a
OFDM Demodulator
656r
OFDM Demodulator
$\{r_1\}$
$\{r_{N_R}\}$
Received Pilot Symbols
160a
RX Spatial Processor
660
Detector
1
N_R
1
N_T
662
Mux
$\{\hat{s}\}$
170a
RX Data Processor
670
Symbol Demapping
672
Channel Deinterleaver
674
Decoder
$\{\hat{d}\}$
Demod Control
Decoding Control
Packet Status
Detector Response
Feedback Info (e.g., Rate, ACK/NAK)
180
Controller
680
Matrix Computation
174
Rate Selector
182
Memory
LUT
682
Channel Estimates
172
Channel Estimator

FIG. 6



N_R Received Symbol Streams
Detected Symbol Stream
Decoded Data Stream
160b
RX Spatial Processor
170b
RX Data Processor
$\{r_1\}$
$\{r_{N_R}\}$
758
Buffer
760
Detector
1
N_R
1
N_T
762
MUX
$\{\hat{s}^n\}$
770
LLR Computation
772
Channel Deinterleaver
$\{x^n\}$
780
Decoder
$\{\hat{d}\}$
782
Channel Interleaver
$\{x^{n+1}\}$
Buffer Control
Detector Response
Demodulation Control
Decoding Control

FIG. 7

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US2004/033680

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L27/26 H04L1/06

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	WO 03/047198 A (QUALCOMM INCORPORATED) 5 June 2003 (2003-06-05) abstract page 3, paragraph 1018 page 7, paragraph 1035 - paragraph 1038 page 8, paragraph 1041 - page 12, paragraph 1058 page 15, paragraph 1073	1-29

-/--

[X] Further documents are listed in the continuation of box C.

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Date of the actual completion of the international search	Date of mailing of the international search report
4 February 2005	17/02/2005

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Palacián Lisa, M

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INTERNATIONAL SEARCH REPORT

International Application No
PCT/US2004/033680

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	KA-WAI NG ET AL: "A simplified bit allocation for V-BLAST based OFDM MIMO systems in frequency selective fading channels" ICC 2002. 2002 IEEE INTERNATIONAL CONFERENCE ON COMMUNICATIONS. CONFERENCE PROCEEDINGS. NEW YORK, NY, APRIL 28 - MAY 2, 2002, IEEE INTERNATIONAL CONFERENCE ON COMMUNICATIONS, NEW YORK, NY : IEEE, US, vol. VOL. 1 OF 5, 28 April 2002 (2002-04-28), pages 411-415, XP010589527 ISBN: 0-7803-7400-2 abstract II. System overview IV. Proposed allocation algorithm	1-29
A	WO 03/075479 A (QUALCOMM INCORPORATED) 12 September 2003 (2003-09-12) page 1, paragraph 1001 page 2, paragraph 1007 - page 3, paragraph 1008 page 6, paragraph 1024 - page 7, paragraph 1031 page 12, line 1044 - page 14, line 1052 page 15, paragraph 1056 - page 16, paragraph 1059 page 19, paragraph 1069 page 21, paragraph 1081 - page 22, paragraph 1082	1-29
A	US 6 141 317 A (MARCHOK DANIEL J 'US! ET AL) 31 October 2000 (2000-10-31) column 2, line 5 - line 27 column 21, line 45 - column 23, line 30	1-29

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INTERNATIONAL SEARCH REPORT

Information on patent family members

International Application No
PCT/US2004/033680

Patent document cited in search report		Publication date	Patent family member(s)			Publication date
WO 03047198	A	05-06-2003	US	2003095506	A1	22-05-2003
			US	2003095508	A1	22-05-2003
			AU	2002346470	A1	10-06-2003
			AU	2002365351	A1	10-06-2003
			EP	1454466	A2	08-09-2004
			EP	1454467	A2	08-09-2004
			WO	03047197	A2	05-06-2003
			WO	03047198	A2	05-06-2003
WO 03075479	A	12-09-2003	US	2003165189	A1	04-09-2003
			AU	2003216479	A1	16-09-2003
			CA	2476061	A1	12-09-2003
			EP	1481489	A1	01-12-2004
			WO	03075479	A1	12-09-2003
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			US	6473394	B1	29-10-2002
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			US	6804192	B1	12-10-2004
			US	6690680	B1	10-02-2004
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			US	2002034160	A1	21-03-2002

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(71) Applicant: TELEBIT CORPORATION [US/US]; 10440 Bubb Road, Cupertino, CA 95014 (US).

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(74) Agent: ALLEN, Kenneth, R.; Townsend and Townsend, One Market Plaza, San Francisco, CA 94105 (US).

(81) Designated States: AT (European patent), AU, BE (European patent), BR, CH (European patent), DE (European patent), DK, FR (European patent), GB (European patent), IT (European patent), JP, KR, LU (European patent), NL (European patent), NO, SE (European patent).

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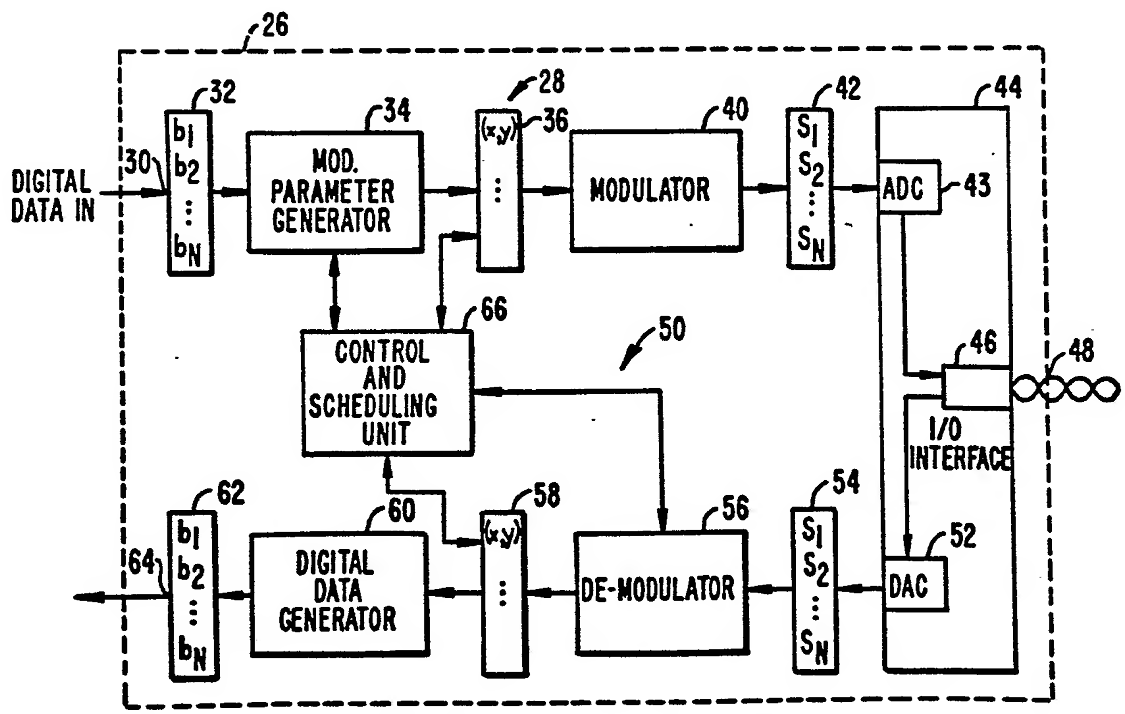
With international search report.

(54) Title: ENSEMBLE MODEM STRUCTURE FOR IMPERFECT TRANSMISSION MEDIA



(57) Abstract

A high speed modem (26) that transmits and receives digital data on an ensemble of carrier frequencies spanning the usable band of a dial-up telephone line (48). The modem includes a system (30, 32, 34, 36, 40, 43, 44) for variably allocating data and power among the carriers to compensate for equivalent noise and to maximize the data rate. Additionally, systems for eliminating the need for an equalization network, for adaptively allocating control of a channel, and for tracking variations in line parameters are disclosed.

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ENSEMBLE MODEM STRUCTURE FOR IMPERFECT TRANSMISSION MEDIA

BACKGROUND OF THE INVENTION

1. Field of the Invention:

The invention relates generally to the field of data communications and, more particularly, to a high speed modem.

2. Description of the Prior Art:

Recently, specially designed telephone lines for the direct transmission of digital data have been introduced. However, the vast majority of telephone lines are designed to carry analog voice frequency (VF) signals. Modems are utilized to modulate VF carrier signals to encode digital information on the VF carrier signals and to demodulate the signals to decode the digital information carried by the signal.

Existing VF telephone lines have several limitations that degrade the performance of modems and limit the rate at which data can be transmitted below desired error rates. These limitations include the presence of frequency dependent noise on the VF telephone lines, a frequency dependent phase delay induced by the VF telephone lines, and frequency dependent signal loss.

Generally, the usable band of a VF telephone line is from slightly above zero to about four kHz. The power spectrum of the line noise is not uniformly distributed over frequency and is generally not determinative. Thus, there is no *a priori* method for determining the distribution of the noise spectrum over the usable bandwidth of the VF line.

Additionally, a frequency-dependent propagation delay is induced by the VF telephone line. Thus, for a complex multi-frequency signal, a phase delay

between the various components of the signal will be induced by the VF telephone line. Again, this phase delay is not determinative and must be measured for an individual VF telephone line at the specific time that transmission takes place.

Further, the signal loss over the VF telephone line varies with frequency. The equivalent noise is the noise spectrum component added to the signal loss component for each carrier frequency, where both components are measured in decibels (dB).

Generally, prior art modems compensate for equivalent line noise and signal loss by gear-shifting the data rate down to achieve a satisfactory error rate. For example, in U.S. patent 4,438,511, by Baran, a high speed modem designated SM9600 Super Modem manufactured by Gandalf Data, Inc., is described. In the presence of noise impairment, the SM9600 will "gear shift" or drop back its transmitted data rate to 4800 bps or 2400 bps. The system described in the Baran patent transmits data over 64 orthogonally modulated carriers. The Baran system compensates for the frequency dependent nature of the noise on the VF line by terminating transmission on carriers having the same frequency as the frequency of large noise components on the line. Thus, Baran gracefully degrades its throughput by ceasing to transmit on carrier frequencies at the highest points of the VF line noise spectrum. The Baran system essentially makes a go/no go decision for each carrier signal, depending on the distribution of the VF line noise spectrum. This application reflects a continuation of the effort initiated by Baran.

Most prior art systems compensate for frequency dependent phase delay induced by the VF line by an equalization system. The largest phase delay is induced in frequency components near the edges of the usable band. Accordingly, the frequency components near the center of the band are delayed to allow the

frequency components at the outside of the band to catch up. Equalization generally requires additional circuitry to accomplish the above-described delays.

A further problem associated with two way transmission over the VF telephone line is that interference between the outgoing and incoming signals is possible. Generally, separation and isolation between the two signals is achieved in one of three ways:

(a) Frequency multiplexing in which different frequencies are used for the different signals. This method is common in modem-based telecommunication systems.

(b) Time multiplexing, in which different time segments are used for the different signals. This method is often used in half-duplex systems in which a transmitter relinquishes a channel only after sending all the data it has. And,

(c) Code multiplexing, in which the signals are sent using orthogonal codes.

All of the above-described systems divide the space available according to constant proportions fixed during the initial system design. These constant proportions, however, may not be suitable to actual traffic load problem presented to each modem. For example, a clerk at a PC work station connected to a remote host computer may type ten or twenty characters and receive a full screen in return. In this case, constant proportions allocating the channel equally between the send and receive modems would greatly overallocate the channel to the PC work station clerk. Accordingly, a modem that allocates channel capacity according to the needs of the actual traffic load situation would greatly increase the efficient utilization of the channel capacity.

SUMMARY OF THE INVENTION

The present invention is a high-speed modem for use with dial-up VF telephone lines. The modem utilizes a multicarrier modulation scheme and variably allocates data and power to the various carriers to maximize the overall data transmission rate. The allocation of power among the carriers is subject to the constraint that the total power allocated must not exceed a specified limit.

In a preferred embodiment, the modem further includes a variable allocation system for sharing control of a communication link between two modems (A and B) according to actual user requirements.

Another aspect of the invention is a system for compensating for frequency dependent phase delay and preventing intersymbol interference that does not require an equalization network.

According to one aspect of the invention, quadrature amplitude modulation (QAM) is utilized to encode data elements of varying complexity on each carrier. The equivalent noise component at each carrier frequency is measured over a communication link between two modems (A and B).

As is known in the art, if the bit error rate (BER) is to be maintained below a specified level, then the power required to transmit a data element of a given complexity on a given carrier frequency must be increased if the equivalent noise component at that frequency increases. Equivalently, to increase data complexity, the signal to noise ratio, S/N, must be increased.

In one embodiment of the present invention, data and power are allocated to maximize the overall data rate within external BER and total available power constraints. The power allocation system computes the marginal required power to increase the symbol rate on each carrier from n to n + 1 information units. The

system then allocates information units to the carrier that requires the least additional power to increase its symbol rate by one information unit. Because the marginal powers are dependent on the values of the equivalent noise spectrum of the particular established transmission link, the allocation of power and data is specifically tailored to compensate for noise over this particular link.

According to another aspect of the invention, a first section of the symbol on each carrier is retransmitted to form a guard-time waveform of duration $T_E + T_{PH}$ where T_E is the duration of the symbol and T_{PH} is the duration of the first section. The magnitude of T_{PH} is greater than or equal to the maximum estimated phase delay for any frequency component of the waveform. For example, if the symbol is represented by the time series, $x_0 \ldots x_{n-1}$, transmitted in time T_E; then the guardtime waveform is represented by the time series, $x_0 \ldots x_{n-1}, x_0 \ldots x_{m-1}$, transmitted in time $T_E + T_{PH}$. The ratio that m bears to n is equal to the ratio that T_{PH} bears to T_E.

At the receiving modem, the time of arrival, T_0, of the first frequency component of the guard-time waveform is determined. A sampling period, of duration T_E, is initiated a time $T_0 + T_{PH}$.

Accordingly, the entire symbol on each carrier frequency is sampled and intersymbol interference is eliminated.

According to a still further aspect of the invention, allocation of control to the transmission link between modems A and B is accomplished by setting limits to the number of packets that each modem may transmit during one transmission cycle. A packet of information comprises the data encoded on the ensemble of carriers comprising one waveform. Each modem is also constrained to transmit a minimum number of packets to maintain the communication link between the modems.

Thus, even if one modem has no data to transmit, the minimum packets maintain timing and other parameters are transmitted. On the other hand, if the volume of data for a modem is large, it is constrained to transmit only the maximum limited number of packets, N, before relinquishing control to the other modem.

In practice, if modem A has a small volume of data and modem B has a large volume of data, modem B will have control of the transmission link most of the time. If control is first allocated to modem A it will only transmit the minimal number, I, of packets. Thus A has control for only a short time. Control is then allocated to B which transmits N packets, where N may be very large. Control is again allocated to modem A which transmits I packets before returning control to B.

Thus, allocation of control is proportional to the ratio of I to N. If the transmission of the volume of data on modem A requires L packets, where L is between I and N, then the allocation is proportional to the ratio of L to N. Accordingly, allocation of the transmission link varies according to the actual needs of the user.

Additionally, the maximum number of packets, N, need not be the same for each modem, but may be varied to accommodate known disproportions in the data to be transmitted by A and B modems.

According to another aspect of the invention, signal loss and frequency offset are measured prior to data determination. A tracking system determines variations from the measured values and compensates for these deviations.

According to a further aspect of the invention, a system for determining a precise value of T_0 is included. This system utilizes two timing signals, at f_1 and f_2, incorporated in a waveform transmitted from modem A at time T_A. The relative phase difference

between the first and second timing signals at time T_A is zero.

The waveform is received at modem B and a rough estimate, T_{EST}, of the time of reception is obtained by detecting energy at f_1. The relative phase difference between the timing signals at time T_{EST} is utilized to obtain a precise timing reference, T_0.

BRIEF DESCRIPTION OF THE DRAWINGS

Fig. 1 is a graph of the ensemble of carrier frequencies utilized in the present invention.

Fig. 2 is a graph of the constellation illustrating the QAM of each carrier.

Fig. 3 is a block diagram of an embodiment of the invention.

Fig. 4 is a flow chart illustrating the synchronization process of the present invention.

Fig. 5 is a series of graphs depicting the constellations for 0, 2, 4, 5, 6 bit data elements and exemplary signal to noise ratios and power levels for each constellation.

Fig. 6 is a graph illustrating the waterfilling algorithm.

Fig. 7 is a histogram illustrating the application of the waterfilling algorithm utilized in the present invention.

Fig. 8 is a graph depicting the effects of phase dependent frequency delay on frequency components in the ensemble.

Fig. 9 is a graph depicting the wave forms utilized in the present invention to prevent intersymbol interference.

Fig. 10 is a graph depicting the method of receiving the transmitted ensemble.

Fig. 11 is a schematic diagram depicting the modulation template.

Fig. 12 is a schematic diagram depicting the quadrants of one square in the modulation template.

Fig. 13 is a schematic diagram of a hardware embodiment of the present invention.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

The present invention is a modem that adaptively allocates power between various carrier frequencies in a frequency ensemble to compensate for frequency dependent line noise, eliminates the need for equalization circuitry to compensate for a frequency dependent phase delay, and provides a duplex mechanism that accounts for varying channel load conditions to allocate the channel between the send and receive modems. Additional features of the invention are described below.

A brief description of the frequency ensemble and modulation scheme utilized in the present invention is first presented with respect to Figs. 1 and 2 to facilitate the understanding of the invention. A specific embodiment of the invention is then described with reference to Fig. 3. Finally, the operation of various features of the invention are described with reference to Figs. 4 through 13.

Modulation and Ensemble Configuration

Referring now to Fig. 1, a diagrammatic representation is shown of the transmit ensemble 10 of the present invention. The ensemble includes 512 carrier frequencies 12 equally spaced across the available 4 kHz VF band. The present invention utilizes quadrature amplitude modulation (QAM) wherein phase independent sine and cosine signals at each carrier frequency are transmitted. The digital information transmitted at a given carrier frequency is encoded by amplitude modulating the independent sine and cosine signals at that frequency.

The QAM system transmits data at an overall bit rate, R_B. However, the transmission rate on each carrier, denoted the symbol or baud rate, R_S, is only a fraction of R_B. For example, if data were allocated equally between two carriers then $R_S = R_B/2$.

In the preferred embodiment 0, 2, 4, 5 or 6 bit data elements are encoded on each carrier and the modulation of each carrier is changed every 136 msec. A theoretical maximum, R_B, assuming a 6 bit R_S for each carrier, of 22,580 bit/sec (bps) results. A typical relizable R_S, assuming 4 bit R_S over 75% of the carriers, is equal to about 11,300 bps. This extremely high R_S is achieved with a bit error rate of less than 1 error/100,000 bits transmitted.

In Fig. 1, a plurality of vertical lines 14 separates each ensemble into time increments known hereafter as "epochs." The epoch is of duration T_E where the magnitude of T_E is determined as set forth below.

The QAM system for encoding digital data onto the various carrier frequencies will now be described with reference to Fig. 2. In Fig. 2 a four bit "constellation" 20 for the nth carrier is depicted. A four bit number may assume sixteen discrete values. Each point in the constellation represents a vector (x_n, y_n) with x_n being the amplitude of the sine signal and y_n being the amplitude of the cosine signal in the above-described QAM system. The subscript n indicates the carrier being modulated. Accordingly, the four bit constellation requires four discrete y_n and four discrete x_n values. As described more fully below, increased power is required to increase the number of bits transmitted at a given carrier frequency due to the equivalent noise component at that frequency. The receive modem, in the case of four bit transmission, must be able to discriminate between four possible values of the x_n and y_n amplitude coefficients. This

ability to discriminate is dependent on the signal to noise ratio for a given carrier frequency.

In a preferred embodiment, packet technology is utilized to reduce the error rate. A packet includes the modulated epoch of carriers and error detection data. Each packet in error is retransmitted until correct. Alternatively, in systems where retransmission of data is undesirable, epochs with forward error correcting codes may be utilized.

Block Diagram

Fig. 3 is a block diagram of an embodiment of the present invention. The description that follows is of an originate modem 26 coupled to an originate end of a communication link formed over a public switched telephone line. It is understood that a communication system also includes an answer modem coupled to the answer end of the communication link. In the following discussion, parts in the answer modem corresponding to identical or similar parts in the originate modem will be designated by the reference number of the originate modem primed.

Referring now to Fig. 3, an incoming data stream is received by a send system 28 of the modem 26 at data input 30. The data is stored as a sequence of data bits in a buffer memory 32. The output of buffer memory 32 is coupled to the input of a modulation parameter generator 34. The output of the modulation parameter generator 34 is coupled to a vector table buffer memory 36 with the vector table buffer memory 36 also coupled to the input of a modulator 40. The output of the modulator 40 is coupled to a time sequence buffer 42 with the time sequence buffer 42 also coupled to the input of a digital-to-analog converter 43 included in an analog I/O interface 44. The interface 44 couples the output of the modem to the public switched telephone lines 48.

A receive system 50 includes an analog-to-digital converter (ADC) 52 coupled to the public switched telephone line 48 and included in the interface 44. The output from the ADC 52 is coupled to a receive time series buffer 54 which is also coupled to the input of a demodulator 56. The output of the demodulator 56 is coupled to a receive vector table buffer 58 which is also coupled to the input of a digital data generator 60. The digital data generator 60 has an output coupled to a receive data bit buffer 62 which is also coupled to an output terminal 64.

A control and scheduling unit 66 is coupled with the modulation parameter generator 34, the vector table buffer 36, the demodulator 56, and the receive vector table buffer 58.

An overview of the functioning of the embodiment depicted in Fig. 3 will now be presented. Prior to the transmission of data, the originate modem 26, in cooperation with the answer modem 26', measures the equivalent noise level at each carrier frequency, determines the number of bits per epoch to be transmitted on each carrier frequency, and allocates power to each carrier frequency as described more fully below.

The incoming data is received at input port 30 and formatted into a bit sequence stored in the input buffer 32.

The modulator 34 encodes a given number of bits into an (x_n, y_n) vector for each carrier frequency utilizing the QAM system described above. For example, if it were determined that four bits were to be transmitted at frequency f_n then four bits from the bit stream would be converted to one of the sixteen points in the four bit constellation of Fig. 2. Each of these constellation points corresponds to one of sixteen possible combinations of four bits. The amplitudes of the sine and cosine signals for frequency n then corresponds to the point in the constellation encoding the four bits

of the bit sequence. The (x_n,y_n) vectors are then stored in the vector buffer table 36. The modulator receives the table of (x_n,y_n) vectors for the carriers in the ensemble and generates a digitally encoded time series representing a wave form comprising the ensemble of QAM carrier frequencies.

In a preferred embodiment the modulator 40 includes a fast Fourier transform (FFT) and performs an inverse FFT operation utilizing the (x,y) vectors as the FFT coefficients. The vector table includes 1,024 independent points representing the 1,024 FFT points of the 512 frequency constellation. The inverse FFT operation generates 1,024 points in a time series representing the QAM ensemble. The 1,024 elements of this digitally encoded time series are stored in the digital time series buffer 42. The digital time sequence is converted to an analog wave form by the analog to digital converter 43 and the interface 46 conditions the signal for transmission over the public switched telephone lines 48.

Turning now to the receive system 50, the received analog waveform from the public switched telephone lines 48 is conditioned by the interface 46 and directed to the analog to digital converter 52. The analog to digital converter 52 converts the analog waveform to a digital 1,024 entry time series table which is stored in the receive time series buffer 54. The demodulator 56 converts the 1,024 entry time series table into a 512 entry (x_n,y_n) vector table stored in the receive vector table buffer 58. This conversion is accomplished by performing an FFT on the time series. Note that information regarding the number of bits encoded onto each frequency carrier has been previously stored in the demodulator and digital data generator 60 so that the (x,y) table stored in the receive vector table buffer 58 may be transformed to an output data bit sequence by the digital data generator 60. For

example, if the (x_n, y_n) vector represents a four bit sequence then this vector would be converted to a four bit sequence and stored in the receive data bit buffer 62 by the digital data generator 60. The receive data bit sequence is then directed to the output 64 as an output data stream.

A full description of the FFT techniques utilized is described in a book by Rabiner et al., entitled Theory and Applications of Digital Signal Processing, Prentice-Hall, Inc., N.J., 1975. However, the FFT modulation technique described above is not an integral part of the present invention. Alternatively, modulation could be accomplished by direct multiplication of the carrier tones as described in the above-referenced Baran patent, which is hereby incorporated by reference, at col. 10, lines 13-70, and col. 11, lines 1-30. Additionally, the demodulation system described in Baran at col. 12, lines 35-70, col. 13, lines 1-70, and col. 14, lines 1-13 could be substituted.

The control and scheduling unit 66 maintains overall supervision of the sequence of operations and controls input and output functions.

Determination of Equivalent Noise

As described above, the information content of the data element encoded on each frequency carrier and the power allocated to that frequency carrier depends on the magnitude of the channel noise component at that carrier frequency. The equivalent transmitted noise component at frequency f_n, $N(f_n)$, is the measured (received) noise power at frequency f_n multiplied by the measured signal loss at frequency f_n. The equivalent noise varies from line to line and also varies on a given line at different times. Accordingly, in the present system, N(f) is measured immediately prior to data transmission.

The steps of a synchronization technique utilized in the present system to measure N(f) and

establish a transmission link between answer and originate modems 26 and 26' are illustrated in Fig. 4. Referring now to Fig. 4, in step 1 the originate modem dials the number of the answer modem and the answer modem goes off hook. In step 2 the answer modem transmits an epoch of two frequencies at the following power levels:

(a) 1437.5 Hz. at -3 dBR; and

(b) 1687.5 Hz at -3 dBR.

The power is measured relative to a reference, R, where, in a preferred embodiment, 0dBR = -9dBm, m being a millivolt. These tones are used to determine timing and frequency offset as detailed subsequently.

The answer modem then transmits an answer comb containing all 512 frequencies at -27dBR. The originate modem receives the answer comb and performs an FFT on the comb. Since the power levels of the 512 frequencies were set at specified values, the control and scheduling unit 66 answer modem 26 compares the (x_n, y_n) values for each frequency of the received code and compares those values to a table of (x_n, y_n) values representing the power levels of the transmitted answer code. This comparison yields the signal loss at each frequency due to the transmission over the VF telephone lines.

During step 3 both the originate and answer modems 26 and 26' accumulate noise data present on the line in the absence of any transmission by either modem. Both modems then perform an FFT on the accumulated noise signals to determine the measured (received) noise spectrum component values at each carrier frequency. Several epochs of noise may be averaged to refine the measurement.

In step 4 the originate modem transmits an epoch of two frequencies followed by an originate comb of 512 frequencies with the same power levels described above for step 2. The answer modem receives the epoch and the originate comb and calculates the timing, fre-

quency offset and signal loss values at each carrier frequency as described above for the originate modem in step 2. At this point the originate modem 26 has accumulated noise and signal loss data for transmission in the answer originate direction while the answer modem has accumulated the same data relating to transmission in the originate answer direction. Each modem requires data relating to transmission loss and receive noise in both the originate-answer and answer-originate directions. Therefore, this data is exchanged between the two modems according to the remaining steps of the synchronization process.

In step 5 the originate modem generates and transmits a first phase encoded signal indicating which carrier frequencies will support two bit transmission at standard power levels in the answer-originate direction. Each component that <u>will</u> support two bits in the answer-originate direction at a standard power level is generated as a -28 dBR signal with 180° relative phase. Each component that <u>will</u> <u>not</u> support two bit transmission in the answer-originate direction at the standard power level is coded as a -28 dBR,0° relative phase signal. The answer modem receives this signal and determines which frequency carriers will support two bit transmission in the answer-originate direction.

In step 6 the answer modem generates and transmits a second phase encoded signal indicating which carrier frequencies will support two bit transmission in both the originate-answer and answer-originate directions. The generation of this signal is possible because the answer modem has accumulated noise and signal loss data in the originate-answer direction and has received the same data for the answer-originate direction in the signal generated by the originate modem in step 5. In the signal generated by the originate modem, each frequency component that <u>will</u> support two bits in <u>both</u> directions is coded with 180° relative

phase and all other components are coded with 0° relative phase.

A transmission link now exists between the two modems. In general, 300 to 400 frequency components will support two bit transmission at a standard power level, thereby establishing about a 600 bit/epoch rate between the two modems. In step 7 the originate modem sends data on the number of bits (0 to 15) and the power levels (0 to 63dB) that can be supported on each frequency in the answer-originate direction in ensemble packets formed over this existing data link. Accordingly, both the originate and answer modem now have the data relating to transmission in the answer-originate direction. The steps for calculating the number of bits and power levels that can be supported on each frequency component will be described below.

In step 8 the answer modem sends data on the number of bits and power levels that can be supported on each frequency in the originate-answer direction utilizing the existing data link. Thus, both modems are apprised of the number of bits and power levels to be supported on each frequency component in both the answer-originate and originate-answer directions.

The above description of the determination of the equivalent noise level component at each carrier frequency sets forth the required steps in a given sequence. However, the sequence of steps is not critical and many of the steps may be done simultaneously or in different order, for example, the performance of the FFT on the originate code and the accumulation of noise data may be done simultaneously. A precise timing reference is also calculated during the synchronization process. The calculation of this timing reference will be described more fully below after the description of the method for calculating the number of bits and power levels allocated to each frequency component.

It is a common VF telephone line impairment that a frequency offset, of up to 7 Hz, exists between transmitted and received signals. This offset must be corrected for the FFT to function reliably. In a preferred embodiment, this correction is achieved by performing a single sideband modulation of the quadrature tones at the offset frequency by the true and Hilbert images of received signal. Synchronization and tracking algorithms generate estimates of the frequency offset necessary.

Power and Code Complexity Allocation

The information encoded on each carrier frequency signal is decoded at the receiver channel by the demodulator 56. Channel noise distorts the transmitted signal and degrades the accuracy of the demodulation process. The transmission of a data element having a specified complexity, e.g., B_0 bits at a specified frequency, f_0, over a VF telephone line characterized by an equivalent noise level component, N_0, will now be analyzed. Generally, external system requirements determine a maximum bit error rate (BER) that can be tolerated. For the transmission of b_0 bits at noise level N_0 and frequency f_0, the signal to noise ratio must exceed E_b/N_0 where E_b is the signal power per bit to maintain the BER below a given BER, $(BER)_0$.

Fig. 5 depicts the QAM constellations for signals of various complexities B. An exemplary signal to noise ratio, E_b/N_0, for each constellation and the power required to transmit the number of bits in the constellation without exceeding $(BER)_0$ is depicted alongside each constellation graph.

A modem operates under the constraint that the total available power placed on the public switched telephone lines may not exceed a value, P_0, set by the telephone companies and government agencies. Thus, signal power may not be increased indefinitely to compensate for line noise. Accordingly, as noise

increases, the complexity of the signals transmitted must be decreased to maintain the required BER.

Most existing modems arbitrarily gear shift the signal complexity down as line noise power increases. For example, one prior art modem reduces the transmitted data rate from a maximum of 9,600 bps to steps of 7,200 bps, 4,800 bps, 2,400 bps, 1,200 bps, and so on until the bit error rate is reduced below a specified maximum. Accordingly, the signal rate is decreased in large steps to compensate for noise. In the Baran patent, the method for reducing the transmission rate takes into account the frequency dependent nature of the noise spectrum. There, each channel carries a preset number of bits at a specified power level. The noise component at each frequency is measured and a decision is made whether to transmit at each carrier frequency. Thus, in Baran, the data rate reduction scheme compensates for the actual distribution of the noise over the available bandwidth.

In the present invention, the complexity of the signal on each frequency carrier and the amount of the available power allocated to each frequency carrier is varied in response to the frequency dependence of the line noise spectrum.

The present system for assigning various code complexities and power levels to the frequency component signals in the ensemble is based on the waterfilling algorithm. The waterfilling algorithm is an information theoretic way of assigning power to a channel to maximize the flow of information across the channel. The channel is of the type characterized by an uneven noise distribution and the transmitter is subject to a power constraint. Fig. 6 provides a visualization of the waterfilling algorithm. Referring now to Fig. 6, power is measured along the vertical axis and frequency is measured along the horizontal axis. The equivalent noise spectrum is represented by the solid line 70 and

the available power is represented by the area of the cross hatched region 72. The name waterfilling comes from the analogy of the equivalent noise function to a series of valleys in a mountain filled with a volume of water representing the assigned power. The water fills the valleys and assumes a level surface. A theoretical description of the waterfilling algorithm is given in the book by Gallagher, entitled Information Theory And Reliable Communication; J. Wiley and Sons, New York, 1968, p. 387.

It must be emphasized that the waterfilling theorem relates to maximizing the theoretical capacity of a channel where the capacity is defined as the maximum of all data rates achievable using different codes, all of which are error correcting, and where the best tend to be of infinite length.

The method utilizing the present invention does not maximize the capacity of the channel. Instead, the method maximizes the amount of information transmitted utilizing the QAM ensemble described above with respect to Fig. 1 and subject to an available power restriction.

An implementation of the waterfilling concept is to allocate an increment of available power to the carrier having the lowest equivalent noise floor until the allocatd power level reaches the equivalent noise level of the second lowest carrier. This allocation requires a scan through the 512 frequencies.

Incremental power is then allocated between the lowest two carriers until the equivalent noise level of the third lowest channel is reached. This allocation level requires many scans through the frequency table and is computationally complex.

The power allocation method used in a preferred embodiment of the present invention is as follows:

(1) Calculate the system noise at the transmitter by measuring the equivalent noise at the receiver and multiplying by transmission loss. This process for measuring these quantities was described above with respect to synchronization and Fig. 4. The system noise components are calculated for each carrier frequency.

(2) For each carrier frequency, calculate the power levels required to transmit data elements of varying complexity (in the present case, 0, 2, 4, 5, 6, and 8 bits). This is accomplished by multiplying the equivalent noise by the signal to noise ratios necessary for transmission of the various data elements with a required BER, for example one error per 100,000 bits. The overall BER is the sum of the signal error rates of each modulated carrier. These signal to noise ratios are available from standard references, and are well-known in the art.

(3) From the calculated required transmission power levels, the marginal required power levels to increase data element complexity are determined. These marginal required power levels are the difference in transmission power divided by the quantitative difference in complexity of the data elements closest in complexity.

(4) For each channel generate a two column table of marginal required power levels and quantitative differences where the units are typically expressed as Watts and bits, respectively.

(5) Construct a histogram by organizing the table of step 4 according to increasing marginal power.

(6) Assign the available transmitter power sequentially over the increasing marginal powers until available power is exhausted.

The power allocation method may be better understood through a simple example. The numbers pre-

sented in the example are not intended to represent parameters encountered in an operating system.

Table 1 sets out the power requirement, P, to transmit a data element of a selected number of bits, N_1, for two carriers A and B at frequencies f_A and f_B.

TABLE 1

Carrier A

N_1	N_2-N_1	P	$MP(N_1 \text{ to } N_2)$
0	-	0	-
2	2	4	MP(0to2)=2/bit
4	2	12	MP(2to4)=4/bit
5	1	19	MP(4to5)=7/bit
6	1	29	MP(5to6)=10/bit

Carrier B

N_1	N_2-N_1	P	$MP(N_1 \text{ to } N_2)$
0	-	0	-
2	2	6	MP(0to2)=3/bit
4	2	18	MP(2to4)=6/bit
5	1	29	MP(4to5)=11/bit
6	1	44	MP(5to6)=15/bit

The marginal power to increase the complexity from a first number of bits, N_1, to a second number of bits, N_2, is defined by the relationship:

$$MP(N_1 \text{ to } N_2) = \frac{P_2 - P_1}{N_2 - N_1}$$

where P_2 and P_1 are the powers required to transmit data elements of complexity N_2 and N_1. N_2-N_1 is quantitative difference in the complexity of the data elements. It is understood the BER is constrained to remain below a preset limit.

The marginal powers for f_A are less than for f_B because the equivalent noise at f_B, $N(f_B)$, is greater than the equivalent noise at f_A, $N(f_A)$.

The implementation of the allocation scheme for carriers A and B will now be described. Assume that a total number of bits, N_T, are encoded on the ensemble but that no bits have been assigned to carriers A or B. For example, $N(f_A)$ and $N(f_A)$ might be greater than the powers of those carriers already carrying the data.

In this example, the system is to allocate ten remaining available power units between carriers A and B to increase the overall data element complexity by the maximum amount.

To increase N_T by two bits requires that four units of power be allocated if channel A is utilized and that six units of power be allocated in channel B is utilized. This follows because for both channels $N_1 = 0$ and $N_2 = 2$ and MP(0 to 2) = 2/bit for channel A and MP(0 to 2) = 3/bit for channel B. Therefore, the system allocates four units of power to carrier A, encodes a two bit data element on carrier A, increases the overall signal complexity from N_T to $N_T + 2$, and has six remaining available power units.

The next increase of two bits requires six power units because MP(2 to 4) = 4/bit for carrier A and MP(0 to 2) = 3/bit for channel B. Therefore, the system allocates six units of power to carrier B, encodes a two bit data element on carrier B, increases the overall signal complexity from $N_T + 2$ to $N_T + 4$ bits, and has no remaining available power units.

As is now clear, the system "shops" among the various carrier frequencies for the lowest power cost to increase the complexity of the overall ensemble data element.

The allocation system is extended to the full 512 carrier ensemble by first generating the tables of

Table 1 for each carrier during a first pass through the frequencies.

A histogram organizing the calculated marginal required power levels for all the carriers according to increasing power is then constructed. Fig. 7 is a depiction of an exemplary histogram constructed according to the present method.

In Fig. 7 the entire table of marginal powers is not displayed. Instead, the histogram is constructed having a range of 64dB with counts spaced in 0.5dB steps. The quantitative differences between the steps are utilized as counts. Although this approach results in a slight round-off error, a significant reduction in task length is achieved. The method used to construct the histogram is not critical to practicing the invention.

Each count of the histogram has an integer entry representing the number of carriers having a marginal power value equal to the power value at the count. The histogram is scanned from the lowest power level. The integer entry at each count is multiplied by the number of counts and subtracted from the available power. The scan continues until available power is exhausted.

When the scan is completed it has been determined that all marginal power values below a given level, MP(max), are acceptable for power and data allocation. Additionally, if available power is exhausted partially through marginal power level, MP(max), then k additional carriers will be allocated power equal to MP(max + 1).

The system then scans through the ensemble again to allocate power and data to the various carriers. The amount of power allocated to each carrier is the sum of marginal power values for that carrier less than or equal to MP(max). Additionally, an amount of power equal to MP(max + 1) will be allocated if the

k MP(max + 1) values have not been previously allocated.

Timing and Phase Delay Compensation

The reconstruction of (x,y) vector table by the receive system requires 1024 time samples of the received waveform. The bandwidth is about 4kHz so that Nyquist sampling rate about 8000/sec and the time sample offset between samples is 125 microseconds. The total sampling time is thus 128 msec. Similarly, the transmit FFT generates a time series having 1024 entries and the symbol time is 128 msec.

The sampling process requires a timing reference to initiate the sampling. This timing reference is established during synchronization by the following method:

During the synchronization steps defined with reference to Fig. 4, the originate modem detects energy at the 1437.5 Hz frequency component (the first timing signal) in the answer comb at time T_{EST}. This time is a rough measure of the precise time that the first timing frequency component arrives at the receiver and is generally accurate to about 2 msec.

This rough measure is refined by the following steps. The first timing signal and a second timing signal (at 1687.5 Hz) are transmitted with zero relative phase at the epoch mark.

The originate modem compares the phases of the first and second timing signals at time T_{EST}. The 250 Hz frequency difference between the first and second timing signals results in an 11° phase shift between the two signals for each 125 microsecond time sample offset. The first and second timing signals have low relative phase distortion (less than 250 microseconds) due to their location near the center of the band. Accordingly, by comparing the phases of the two timing samples and correcting T_{EST} by the number of

time sampling offsets indicated by the phase difference, a precise timing reference, T_0, can be determined.

A further difficulty relating to timing the sampling process relates to frequency dependent phase delay induced by the VF line. This phase delay typically is on the order of 2 msec, or more, for VF telephone lines. Further, this phase delay is significantly worse near the edges of the 4kHz usable band.

Fig. 8 depicts distribution of the frequency carriers of the ensemble after undergoing frequency dependent phase delay. Referring to Fig. 8, three signals 90, 92, and 94 at frequencies f_0, f_{256}, and f_{512} are depicted. Two symbols, x_i and y_i, of length T_S are transmitted at each frequency. Note that the duration of each symbol is not changed. However, the leading edge of the signals near the edge of the band 92 and 94 are delayed relative to those signals near the center of the band 94.

Additionally, for two sequentially transmitted epochs x_i and y_i the trailing section of the first symbol x_i on signals 92 and 96, near the outer edge of the band will overlap the leading edge of the second symbol y_i on the signal 94 near the center of the band. This overlap results in intersymbol interference.

If the sampling interval is framed to sample a given time interval, T_S, then complete samples of every carrier in the ensemble will not be obtained and signals from other epochs will also be sampled.

Existing systems utilize phase correction (equalization) networks to correct for phase distortion and to prevent intersymbol interference.

The present invention utilizes a unique guard-time format to eliminate the need for an equalization network. This format is illustrated in Fig. 9.

Referring now to Fig. 9, first, second, and third transmitted symbols, represented by time series x_i, y_i, and z_i, respectively, are depicted. The wave-

forms depicted in Fig. 3 are modulated on one of the carriers at frequency f. In this example a symbol time, T_S, of 128 msec. and a maximum phase delay, T_{PH}, of 8 msec are assumed. A guard-time waveform is formed by repeating the first 8 msec. of the symbol. The guard-time waveform defines an epoch of 136 msec. For example, in the first waveform 110, (X_i), the time series of the symbol, $X_0 - X_{1023}$, is first transmitted, then the first 8 msec. of the symbol, $X_0 - X_{63}$, are repeated.

The sampling of the epoch is aligned with the <u>last</u> 128 msec. of the guard-time waveform (relative to the beginning of the guard-time epoch defined by those frequency components which arrive first).

This detection process is illustrated in Fig. 10. In Fig. 10 first and second guard-time waveforms 110 and 112 at f_1, near the center of the band, and f_2, near the edge of the band, are depicted. The frequency component at f_1 is the component of the ensemble that arrives first at the receiver and the component at f_2 arrives last. In Fig. 10 the second waveform 112, at f_2, arrives at the receiver at $T_0 + T_{PH}$, which is 8 msec. after the time, T_0, that the first waveform 110, at f_1, arrives at the receiver. The sampling period of 128 msec. is initiated at the time $T_0 + T_{PH}$. Thus, the entire symbol on f_2, $X_0 - X_{1023}$, is sampled. The entire symbol at f_1 is also sampled because the initial 8 msec. of that symbol has been retransmitted.

Also, intersymbol interference has been eliminated. The arrival of the second symbol, (y_i), at f_1 has been delayed 8 msec. by the retransmission of the first 8 msec. of (x_i). Thus, the leading edge of the second symbol at f_1, does not overlap the trailing edge of the first symbol at f_2.

The 8 msec. guardtime reduces the usable time-bandwidth product of the system by only about 6%. This

small decrease is due to the very long duration of each symbol relative to the necessary guardtime.

Tracking

In practice, for a given carrier, the magnitudes of the (x,y) vectors extracted during the demodulation process do not fall exactly at the constellation points but are distributed over a range about each point due to noise and other factors. Accordingly, the signal is decoded utilizing a modulation template as depicted in Fig. 11.

Referring now to Fig. 11, the template is formed by a grid of squares 113 with the constellation points 114 at the centers of the squares 113.

In Fig. 11, the vector $W = (x_n, y_n)$ represents the demodulated amplitudes of the sine and cosine signals at f_n. W is in the square 113 having the constellation point (3,3) centered therein. Accordingly, W is decoded as (3,3).

The present invention includes a system for tracking to determine changes in transmission loss, frequency offset, and timing from the values determined during synchronization.

This tracking system utilizes the position of the received vectors in the squares of the demodulation template of Fig. 11. In Fig. 12, a single square is divided into four quadrants upper left, lower right, upper right, lower right, 115, 116, 117, and 118 characterized as too fast, too slow, too big, and too little, respectively. If counts in all four quadrants over time by frequency or over frequency at one time are equal or nearly equal then the system is in alignment. That is, if noise is the only impairment, then the direction of error for the decoded vector, W, should be random.

However, if transmission loss changes by even 0.1dB the number of too small counts will vary significantly from the number of too large counts. Similarly, a large difference between the number of too fast and

too slow counts indicates a phase rotation caused by a change in the offset frequency. Thus, the differences between the too fast, too slow, and too big, too small counts is an error characteristic that tracks variations in signal loss and offset frequency.

The present invention utilizes this error characteristic to adjust the signal loss and frequency offset determined during synchronization. For each frequency an adjustment of ± .1dB or ± 1.0° is made depending on the error characteristic. Other divisions of the decoding region into distinct or overlapping subregions characterized as too fast, too slow, too big, and too little are preferred in some embodiments.

Additionally, the phase of the timing signals is tracked to allow corrections of T_0.

Allocation of Channel Control

The present invention further includes a unique system for allocating control of an established communication link between the originate and answer modems (hereinafter designated A and B, respectively). Each waveform comprising the encoded ensemble of frequencies forms a packet of information.

Control of the transmission link is first allocated to modem A. Modem A then determines the volume of data in its input buffer and transmits between I (a minimum) and N (a previously determined maximum) packets of data as appropriate. The predetermined number N serves as a limit and the end number of transmitted packets may be significantly less than required to empty the input buffer. On the other hand, if modem A has little or no data in its input buffer it will still transmit I packets of information to maintain communication with modem B. For example, the I packets may comprise the originate or answer comb of frequencies defined above with respect to Fig. 4 and the synchronization process.

Control of the communication link is then allocated to modem B which repeats the actions of modem A. Of course, if modem B transmits the minimum number, I, of packets it is confirming to modem A the vitality of modem B.

There is no need for the limits N on the two modems to be the same, or to restrict them from being adaptable under modem control to obtain rapid character echo or other user oriented goals.

Hardware Implementation

Fig. 13 is a block diagram of a hardware embodiment of the invention. Referring now to Fig. 13, an electronic digital processor 120, an analog I/O interface 44, and a digital I/O interface 122 are coupled to a common data bus 124. The analog I/O interface 44 interfaces the public switched telephone line 48 with the common data bus 124 and the digital interface 122 interfaces digital terminal equipment 126 with the common data bus 124.

The following components are utilized in a preferred embodiment of the invention. The analog I/O interface 44 is a high performance 12 bit coder-decoder (codec) and telephone line interface. The interface has access to RAM 132 and is controlled by supervisory microprocessor 128. The codec is a single chip combination of an analog to digital converter, a digital to analog converter, and several band pass filters.

The digital I/O interface 122 is a standard RS-232 serial interface to a standard twenty-five pin RS-232 type connector or a parallel interface to a personal computer bus.

The electronic digital processor 120, includes a supervisory processor 128, a general purpose mathematical processor 130, a 32K by 16 bit shared RAM subsystem 132, and a read only memory (ROM) unit 133, coupled to an address bus 135.

The supervisory microprocessor 128 is a 68000 data processor subsystem including a 10MHz 68000 processor and the 68000 program memory. The 32K by 16 bit program memory consists of several low power, high density, ROM chips included in the ROM unit 133.

The mathematical processor 130 is a 320 digital signal microprocessor system (DSP) including a 20MHz 320 processor, the 320 program memory, and an interface to the shared RAM system. Two high speed ROM chips, included in ROM unit 133, comprise the 8192 x 16 bit program memory.

The 320 system program memory includes programs for performing the modulation table look-up, FFT, demodulation, and other operations described above. The 68000 processor handles digital data streams at the input and output, performs tasking to and supervision of the 320 signal processor and associated analog I/O, and performs self and system test as appropriate.

The invention has been explained with respect to specific embodiments. Other embodiments will now be apparent to those of ordinary skill in the art.

In particular, the ensemble of carrier frequencies need not be limited as above-described. The number of carriers may be any power of 2, e.g. 1024, or some arbitrary number. Additionally, the frequencies need not be evenly spaced over the entire VF band. Further, the QAM scheme is not critical to practicing the invention. For example, AM could be utilized although the data rate, R_B, would be reduced.

Still further, the modulation template need not be comprised of squares. Arbitrarily shaped regions surrounding the constellation points may be defined. The tracking system was described where the squares in the modulation template were divided into four quadrants. However, a given parameter may be tracked by tracking the difference in the number of

counts in arbitrary regions defined about a constellation point.

Still further, a hardware embodiment including a supervisory microprocessor and a general purpose mathematical processor has been described. However, different combinations of IC chips may be utilized. For example, a dedicated FFT chip could be utilized to perform modulation and demodulation operations.

Still further, the information units utilized in the above description were bits. However, the invention is not limited to binary system.

Accordingly, it is therefore intended that the invention can be limited except as indicated by the appended claims.

WHAT IS CLAIMED IS:

1. In a high speed modem, for transmitting data over a telephone line, of the type that encodes data elements on an ensemble of carrier frequencies, a method for allocating data and power to the carrier frequencies, said method comprising the steps of:

determining the equivalent noise component for every carrier frequency in the ensemble;

determining the marginal power requirement to increase the complexity of the data element on each carrier from n information units to n + 1 information units, n being an integer between 0 and N;

ordering the marginal powers of all the carriers in the ensemble in order of increasing power;

assigning available power to the ordered marginal powers in order of increasing power;

determining the value, MP(max) at which point the available power is exhausted; and

allocating power and data to each carrier frequency where the power allocated is equal to the sum of all the marginal powers less than or equal to MP(max) for that carrier and the number of data units allocated is equal to the number of marginal powers for that carrier less than or equal to MP(max).

2. The invention of claim 1 where said step of ordering comprises the steps of:

providing a table of arbitrary marginal power levels; and

rounding the value of each determined marginal power level to one of the values of the table of arbitrary marginal power levels to decrease computational complexity.

3. The invention of claim 2 wherein the step of determining equivalent noise comprises the steps of:

providing an A and a B modem interconnected by a telephone line;

establishing a communication link between said A and B modems;

accumulating line noise data during a no transmission time interval at said A and B modems;

transmitting at least a first ensemble of frequency carriers from said A modem to said B modem, where the amplitude of each carrier has a predetermined value;

receiving said first ensemble at said B modem;

measuring the amplitude of each carrier received at said B modem;

comparing the measured amplitudes at said B modem with said predetermined amplitudes to determine signal loss, in dB, at each carrier frequency;

determining the value of the component, in dB, at each carrier frequency of the accumulated noise; and

adding the signal loss at each carrier frequency to the noise component at each carrier frequency to determine equivalent noise.

4. A high speed modem of the type for transmitting a signal on a VF telephone line, comprising:

means for receiving an input digital data stream and for storing said input digital data;

means for generating a modulated ensemble of carriers to encode said input digital data, where each carrier has data elements of variable complexity encoded thereon;

means for measuring the signal loss and noise loss of the VF telephone line for each carrier; and

means for varying the complexity of the data element encoded on each carrier and the amount of power allocated to each carrier to compensate for the measured signal loss and noise level.

5. A high speed modem of the type that encodes data elements on an ensemble of carriers of different frequency, said modem comprising:

a digital electronic processor;

a digital electronic memory;

bus means for coupling said processor and said memory;

means, associated with said digital electronic processor, for

determining the equivalent noise component for every carrier frequency in the ensemble;

determining the marginal power requirements to increase the complexity of the data element on each carrier from n information units to n + 1 information units, n being an integer between 0 and N;

ordering the marginal powers of all the carriers in the ensemble in order of increasing power;

assigning available power to the ordered marginal powers in order of increasing power;

determining the value, MP(max) at which point the available power is exhausted; and

assigning power and data to each carrier frequency where the power assigned is equal to the sum of all the marginal powers less than or equal to MP(max) for that carrier and the number of data units is equal to the number of marginal powers for that carrier less than or equal to MP(max).

6. In a high speed modem, for transmitting data in the form of a QAM ensemble of carrier frequencies on a VF telephone line, of the type that measures the magnitude of a system parameter prior to

transmission, a method for tracking deviations in the magnitude of the system parameter during the receipt of data, said method comprising the steps of:

generating QAM constellations for a plurality of carrier frequencies;

constructing a demodulation template for one of said plurality of carrier frequencies comprising a plurality of first regions with one of the points of said constellation positioned within each of said first regions;

forming a set of tracking regions where each first region has a first and second tracking region disposed therein;

demodulating said ensemble of carriers to obtain the demodulation points positioned in said set of first and second tracking regions;

counting the number of points disposed in said set of first tracking regions and the number of points disposed in said set of second tracking regions;

determining the difference in the number of counts disposed in said set of first tracking regions and disposed in said tracking regions to construct an error characteristic; and

utilizing said error characteristic to adjust the magnitude of said signal parameter during the receipt of data.

7. The invention of claim 6 wherein said step of constructing a demodulation template comprises the step of:

constraining said first regions to be in the shape of squares having said constellation points centered therein.

8. The invention of claim 7 wherein said step of forming said tracking regions comprises the step of:

dividing said squares into quadrants; and

selecting said tracking regions to be symmetrically disposed quadrants.

9. In a communication system of the type including two modems (A and B) coupled by a transmission link, each modem having an input buffer for storing data to be transmitted, a method for allocating control of the transmission link between modem A and B comprising the steps of:

allocating control of the transmission link to modem A;

determining the volume of data stored in the input buffer of modem A;

determining the number,K, of packets of data required to transmit the volume of data stored in the input buffer of modem A;

transmitting L packets of data from modem A to modem B where L is equal to I_A if K is less than I_A, where L is equal to K if K is greater than or equal to I_A, and where L is equal to N_A if K is greater than N_A so that the minimum number of packets transmitted is I_A and the maximum is N_A;

allocating control of the transmission link to modem B;

determining the volume of data in the input buffer of modem B;

determining the number,J, of packets of data required to transmit the volume of data stored in the input buffer of modem B;

transmitting M packets of data from modem B to modem A where M is equal to I_B if J is less than I_B, where M is equal to J if J is greater than or equal to I_B, and where L is equal to N_B if J is greater than N_B so that the minimum number of packets transmitted is I_B and the maximum is N_B;

where allocation of control between modem A and B is dependent on the volume of data stored in the input buffers of modems A and B.

10. In a high speed modem, for transmitting data over a telephone line, of the type that encodes data elements on an ensemble of carrier frequencies, a system for allocating data and power to the carrier frequencies, said system comprising:

means for determining the equivalent noise component for every carrier frequency in the ensemble;

means for determining the marginal power requirement to increase the complexity of the data element on each carrier from n information units to n + 1 information units, n being an integer between 0 and N;

means for ordering the marginal powers of all the carriers in the ensemble in order of increasing power;

means for assigning available power to the ordered marginal powers in order of increasing power;

means for determining the value, MP(max) at which point the available power is exhausted; and

means allocating power and data to each carrier frequency where the power allocated is equal to the sum of all the marginal powers less than or equal to MP(max) for that carrier and the number of data units allocated is equal to the number of marginal powers for that carrier less than or equal to MP(max).

11. The invention of claim 10 where said means for ordering comprises:

means for providing a table of arbitrary marginal power levels; and

means for rounding the value of each determined marginal power level to one of the values of the table of arbitrary marginal power levels to decrease computational complexity.

12. The invention of claim 11 wherein an A and B modem are connected by a telephone line and the means for determining equivalent noise comprises:

means for establishing a communication link between said A and B modems;

means for accumulating line noise data during a no transmission time interval at said A and B modems;

means for transmitting a first ensemble of frequency carriers from said A modem to said B modem, where the amplitude of each carrier has a predetermined value;

means for receiving said first ensemble at said B modem;

means for measuring the amplitude of each carrier received at said B modem;

means for comparing the measured amplitudes at said B modem with said predetermined amplitudes to determine signal loss at each carrier frequency;

means for determining the value of the component, in dB, at each carrier frequency of the accumulated noise; and

means for adding the signal loss at each carrier frequency to the noise component at each carrier frequency to determine equivalent noise.

13. In a high speed modem, for transmitting data in the form of a QAM ensemble of carrier frequencies on a VF telephone line, of the type that measures the magnitude of a system parameter prior to transmission, a system for tracking deviations in the magnitude of the system parameter during the receipt of data, said system comprising:

means for generating QAM constellations for a plurality of carrier frequencies;

means for constructing a demodulation template for one of said plurality of carrier frequencies comprising a plurality of first regions with one of the

points of said constellation positioned within each of said first regions;

means for forming a set of tracking regions where each first region has a first and second tracking region disposed therein;

means for demodulating said ensemble of carriers to obtain the modulation points positioned in said set of first and second tracking regions;

means for counting the number of points disposed in said set of first tracking regions and the number of points disposed in said set of second tracking regions;

means for determining the difference in the number of counts disposed in said set of first tracking regions and disposed in said tracking regions to construct an error characteristic; and

means for utilizing said error characteristic to adjust the magnitude of said signal parameter during the receipt of data.

14. The invention of claim 13 wherein said means for constructing a demodulation template comprises:

means for constraining said first regions to be in the shape of squares having said constellation points centered therein.

15. The invention of claim 14 wherein said means for forming said tracking regions comprises:

means for dividing said squares into quadrants; and

means for selecting said tracking regions to be symmetrically disposed quadrants.

16. In a communication system of the type including two modems (A and B) coupled by a transmission link, each modem having an input buffer for storing data to be transmitted, a system for allocating control

of the transmission link between modem A and B comprising:

means for allocating control of the transmission link to modem A;

means for determining the number, K, of packets of data required to transmit the volume of data stored in the input buffer of modem A;

means for transmitting L packets of data from modem A to modem B where L is equal to I_A if K is less than I_A but less than N_A, where L is equal to K if K is greater than or equal to I_A, and where L is equal to N_A if K is greater than N_A so that the minimum number of packets transmitted is I_A and the maximum is N_A;

means for allocating control of the transmission link to modem B;

means for determining the volume of data in the input buffer of modem B;

means for determining the number, J, of packets of data required to transmit the volume of data stored in the input buffer of modem B;

means for transmitting M packets of data from modem B to modem A where M is equal to I_B if J is less than I_B, where M is equal to J if J is greater than or equal to I_B but less than N_B, and where L is equal to N_B if J is greater than N_B so that the minimum number of packets transmitted is I_B and the maximum is N_B;

where allocation of control between modem A and B is dependent on the volume of data stored in the input buffers of modems A and B.

17. In a high speed modem communication system including two modems (A and B) coupled by a transmission link, each modem having an input buffer for storing data to be transmitted, each modem for transmitting data over a telephone line and each modem of the type that encodes data elements on an ensemble of carrier frequencies, a method of operating said modems to effi-

ciently allocate power and data to the carrier frequencies, to compensate for frequency dependent phase delay, where the maximum estimated magnitude of the phase delay is T_{PH}, to prevent intersymbol interference, to allocate control of the transmission link between modem A and modem B and for initiating a sampling interval having a given time sample offset equal to the reciprocal of the sampling frequency, said method comprising:

determining the equivalent noise component for every carrier frequency in the ensemble;

determining the marginal power requirement to increase the complexity of the data element on each carrier from n information units to n + 1 information units, n being an integer between 0 and N;

ordering the marginal powers of all the carriers in the ensemble in order of increasing power;

assigning available power to the ordered marginal powers in order of increasing power;

determining the value, MP(max) at which point the available power is exhausted;

allocating power and data to each carrier frequency where the power allocated is equal to the sum of all the marginal powers less than or equal to MP(max) for that carrier and the number of data units allocated is equal to the number of marginal powers for that carrier less than or equal to MP(max);

transmitting a symbol encoded on one of said carrier frequencies where said symbol is a predetermined time duration, T_S;

retransmitting the first T_{PH} seconds of said symbol to form a transmitted waveform of duration $T_E + T_{PH}$;

allocating control of the transmission link to modem A;

determining the volume of data stored in the input buffer of modem A;

determining the number, K, of packets of data required to transmit the volume of data stored in the input buffer of modem A;

transmitting L packets of data from modem A to modem B where L is equal to I_A if K is less than I_A, where L is equal to K if K is greater than or equal to I_A, and where L is equal to N_A if K is greater than N_A so that the minimum number of packets transmitted is I_A and the maximum is N_A;

allocating control of the transmission link to modem B;

determining the volume of data in the input buffer of modem B;

determining the number,J, of packets of data required to transmit the volume of data stored in the input buffer of modem B;

transmitting M packets of data from modem B to modem A where M is equal to I_B if J is less than I_B, where M is equal to J if J is greater than or equal to I_B, and where L is equal to N_B if J is greater than N_B so that the minimum number of packets transmitted is I_B and the maximum is N_B;

where allocation of control between modem A and B is dependent on the volume of data stored in the input buffers of modems A and B;

generating an analog waveform at modem A including first and second frequency components at f_1 and f_2;

transmitting said waveform from modem A to modem B at time T_A;

adjusting the phases of said first and and second frequency components so that their relative phase difference at time T_A is equal to about 0°;

detecting energy at frequency f_1 at modem B to determine the estimated time, T_{EST}, that said waveform arrives at modem B;

determining the relative phase difference at modem B between said first and second frequency components at time T_{EST};

calculating the number of sampling time offsets, N_I, required for the relative phase of said first and second carriers to change from 0 to said relative phase difference; and

changing the magnitude of T_{EST} by N_I sampling intervals to obtain a precise timing reference, T_0.

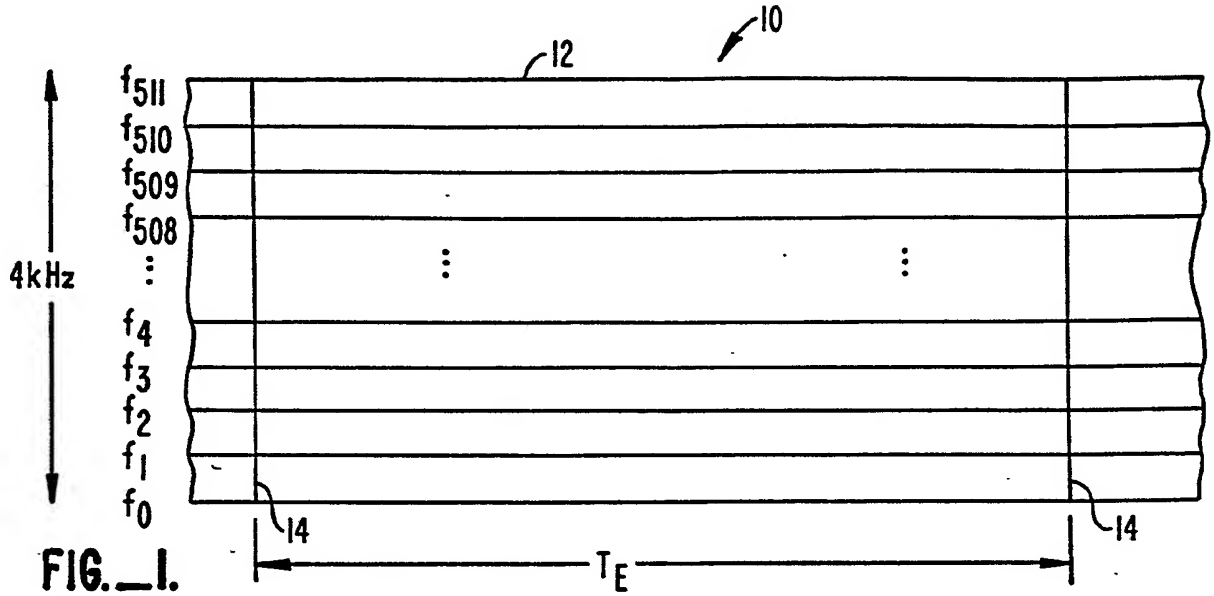
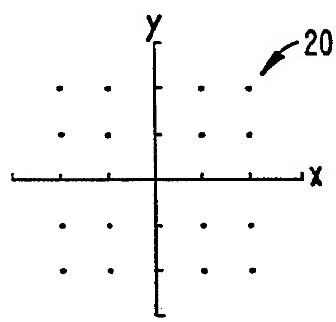


FIG._1.

FIG._2.



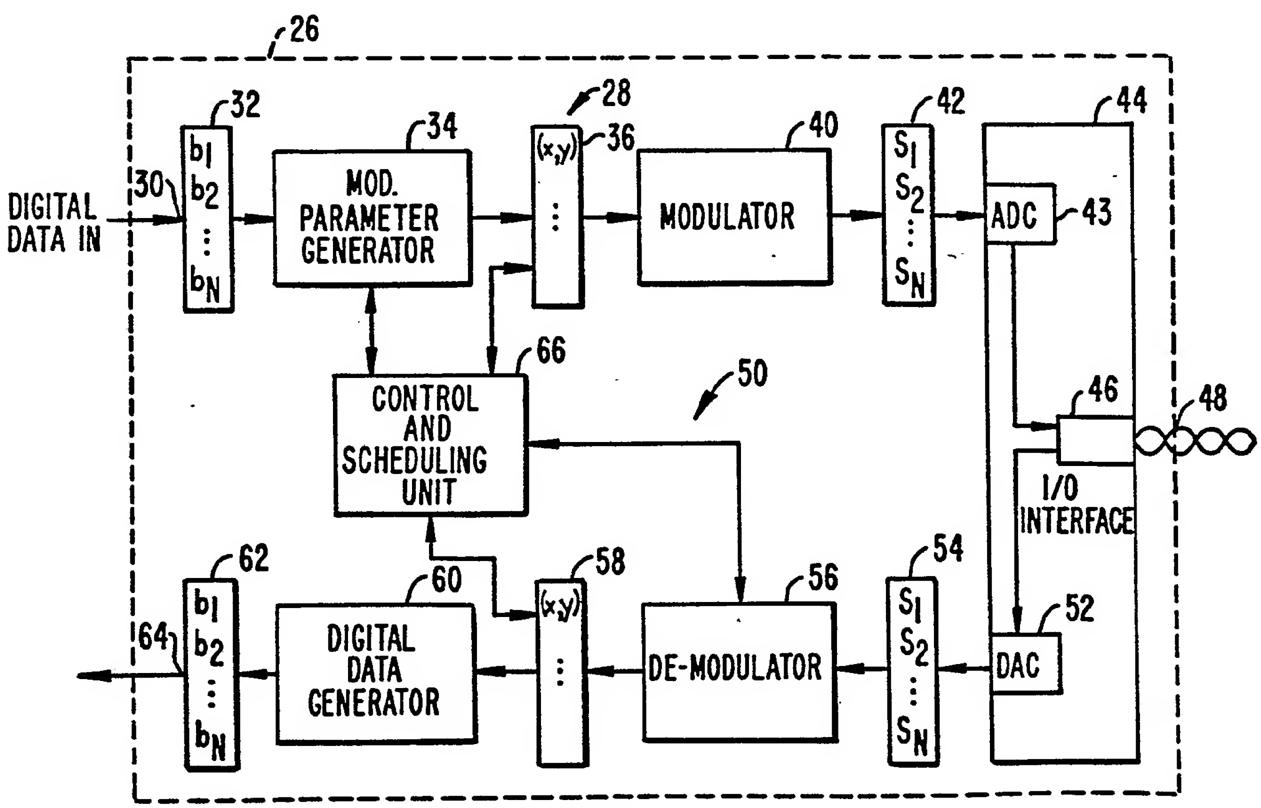


FIG._3.

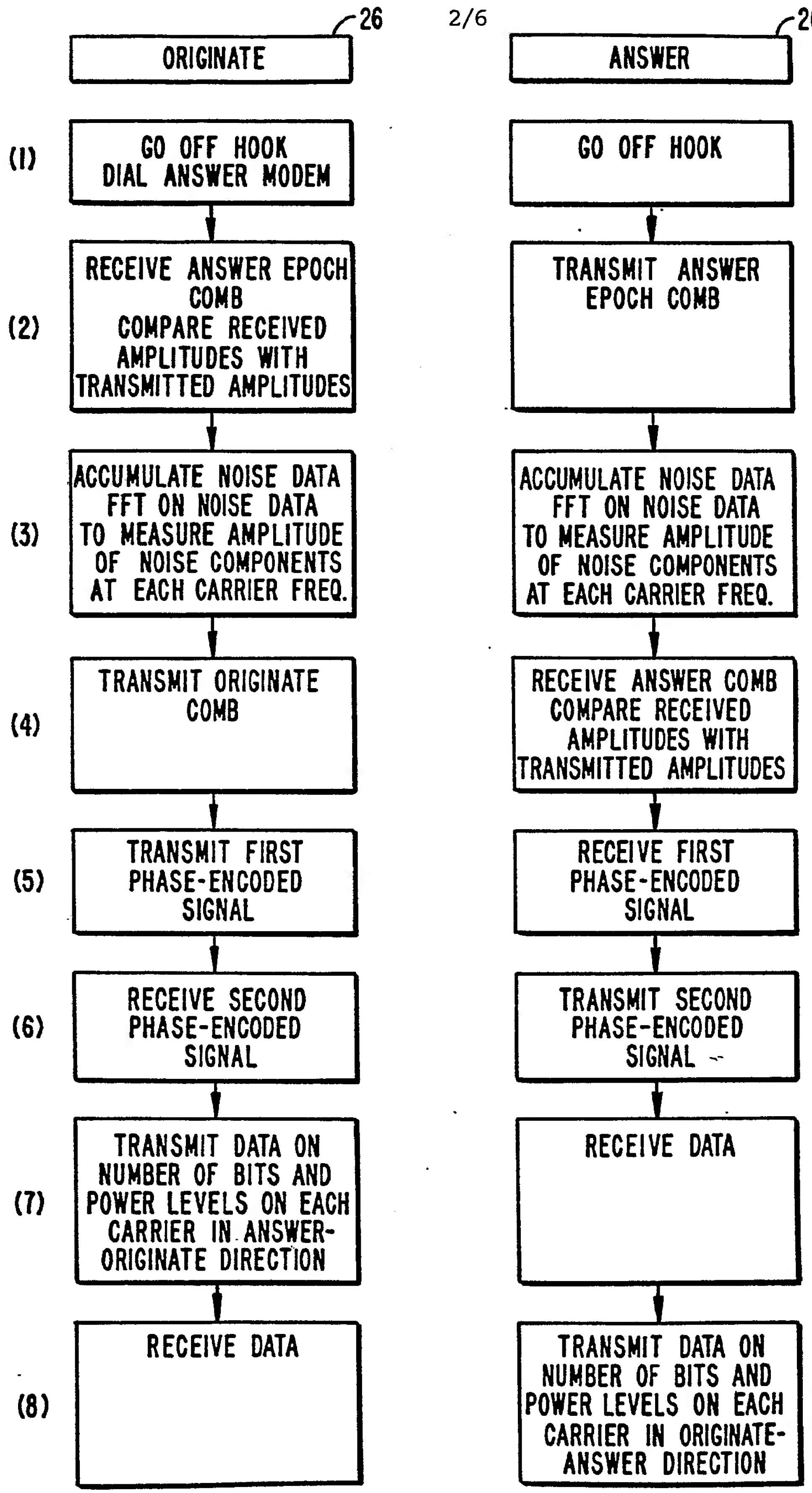
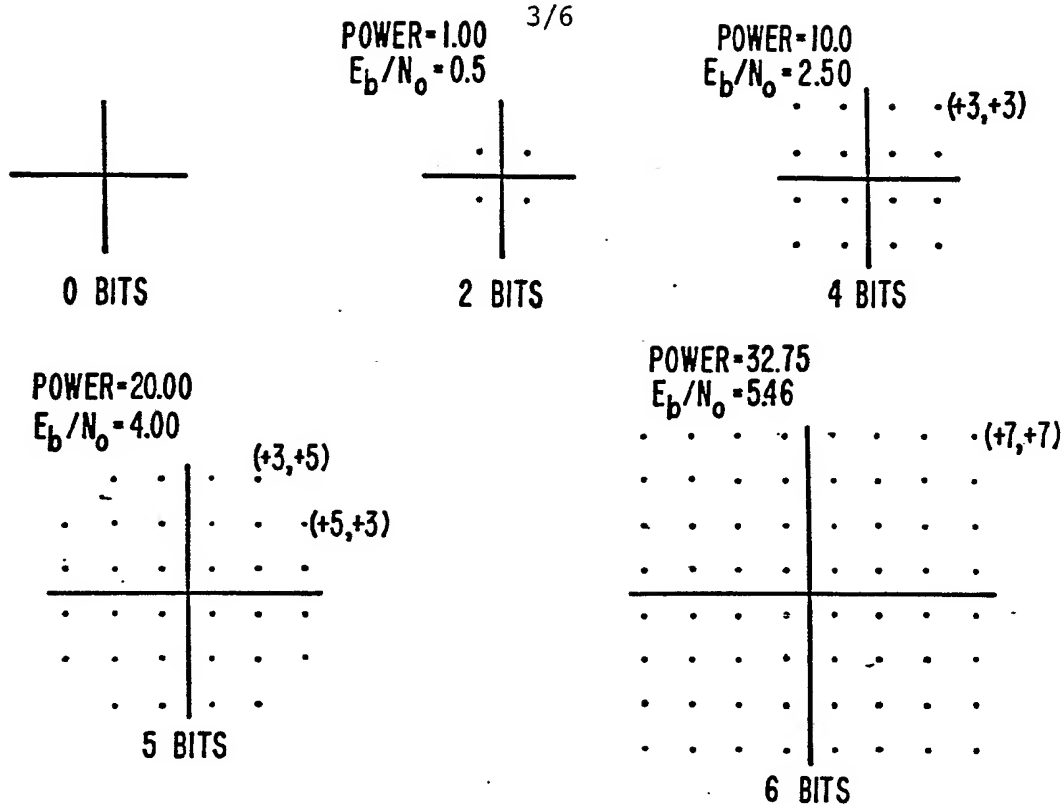
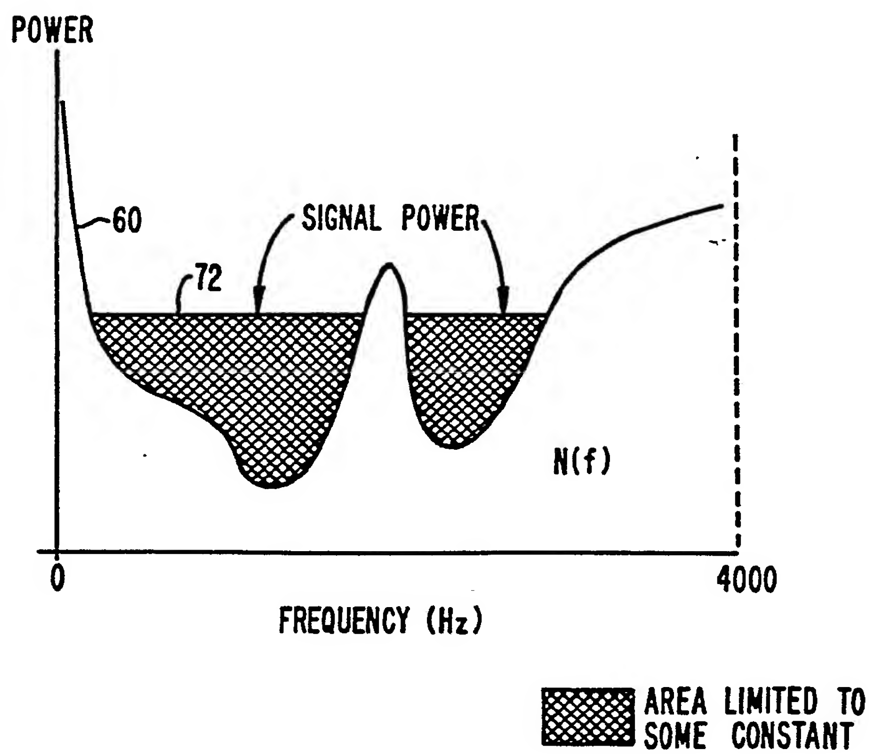

26
ORIGINATE
26'
ANSWER
(1)
GO OFF HOOK DIAL ANSWER MODEM
GO OFF HOOK
(2)
RECEIVE ANSWER EPOCH COMB COMPARE RECEIVED AMPLITUDES WITH TRANSMITTED AMPLITUDES
TRANSMIT ANSWER EPOCH COMB
(3)
ACCUMULATE NOISE DATA FFT ON NOISE DATA TO MEASURE AMPLITUDE OF NOISE COMPONENTS AT EACH CARRIER FREQ.
ACCUMULATE NOISE DATA FFT ON NOISE DATA TO MEASURE AMPLITUDE OF NOISE COMPONENTS AT EACH CARRIER FREQ.
(4)
TRANSMIT ORIGINATE COMB
RECEIVE ANSWER COMB COMPARE RECEIVED AMPLITUDES WITH TRANSMITTED AMPLITUDES
(5)
TRANSMIT FIRST PHASE-ENCODED SIGNAL
RECEIVE FIRST PHASE-ENCODED SIGNAL
(6)
RECEIVE SECOND PHASE-ENCODED SIGNAL
TRANSMIT SECOND PHASE-ENCODED SIGNAL
(7)
TRANSMIT DATA ON NUMBER OF BITS AND POWER LEVELS ON EACH CARRIER IN ANSWER-ORIGINATE DIRECTION
RECEIVE DATA
(8)
RECEIVE DATA
TRANSMIT DATA ON NUMBER OF BITS AND POWER LEVELS ON EACH CARRIER IN ORIGINATE-ANSWER DIRECTION

FIG.—4.



POWER=1.00
E_b/N_o=0.5
POWER=10.0
E_b/N_o=2.50
(+3,+3)
0 BITS
2 BITS
4 BITS
POWER=20.00
E_b/N_o=4.00
(+3,+5)
(+5,+3)
5 BITS
POWER=32.75
E_b/N_o=5.46
(+7,+7)
6 BITS

FIG._5.



POWER
60
72
SIGNAL POWER
N(f)
0
4000
FREQUENCY (Hz)
AREA LIMITED TO SOME CONSTANT

FIG._6.

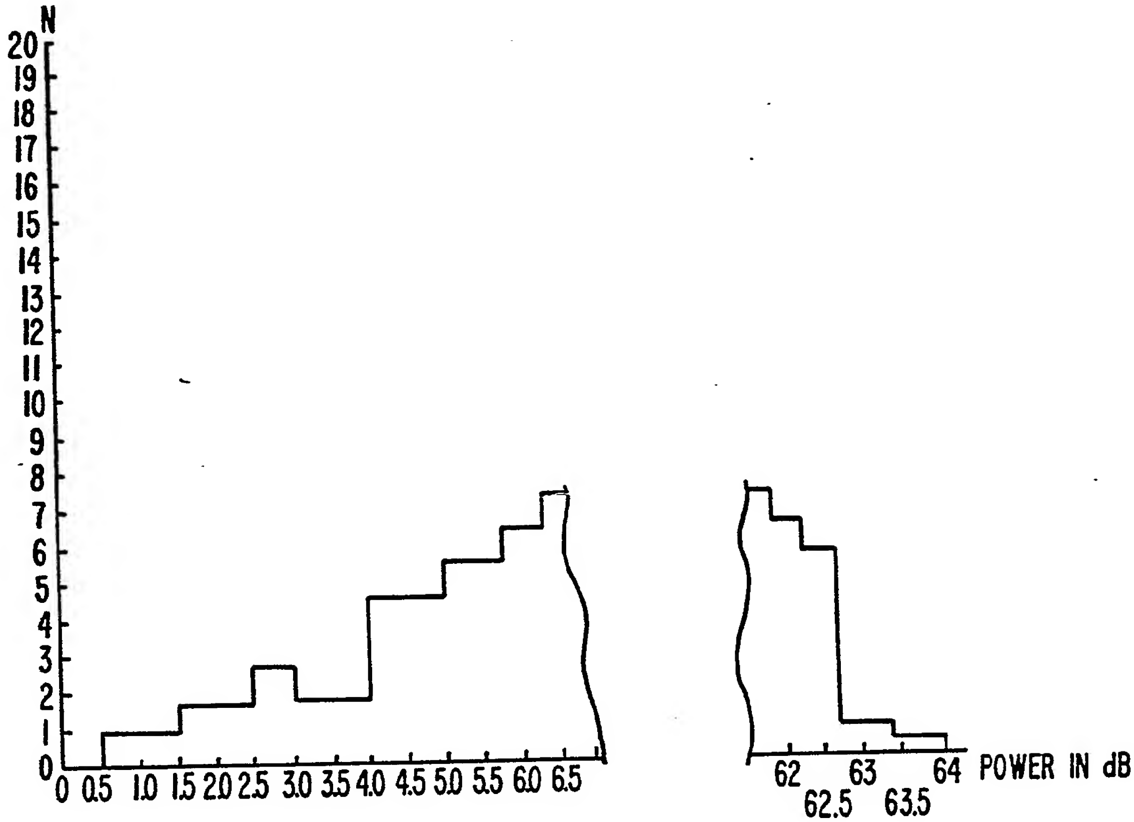

N
20
19
18
17
16
15
14
13
12
11
10
9
8
7
6
5
4
3
2
1
0
0 0.5 1.0 1.5 2.0 2.5 3.0 3.5 4.0 4.5 5.0 5.5 6.0 6.5
62 62.5 63 63.5 64
POWER IN dB

FIG._7.

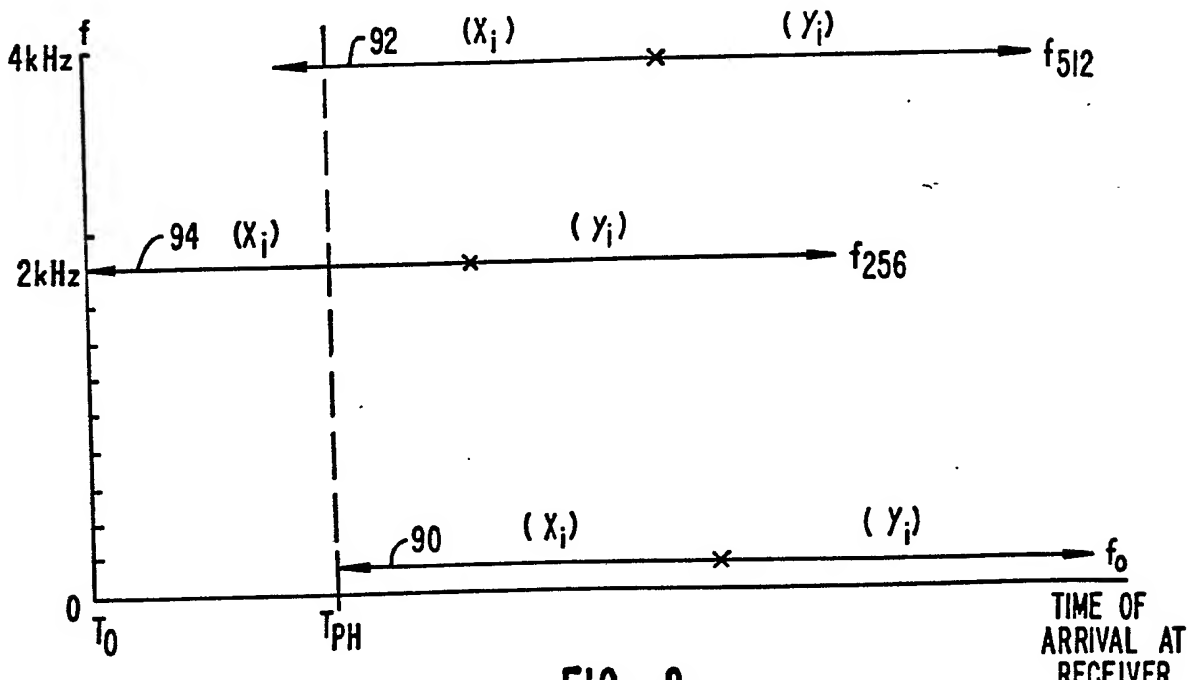

f
4kHz
2kHz
0
92
94
90
(Xi)
(Yi)
f512
f256
f0
T0
TPH
TIME OF ARRIVAL AT RECEIVER

FIG._8.

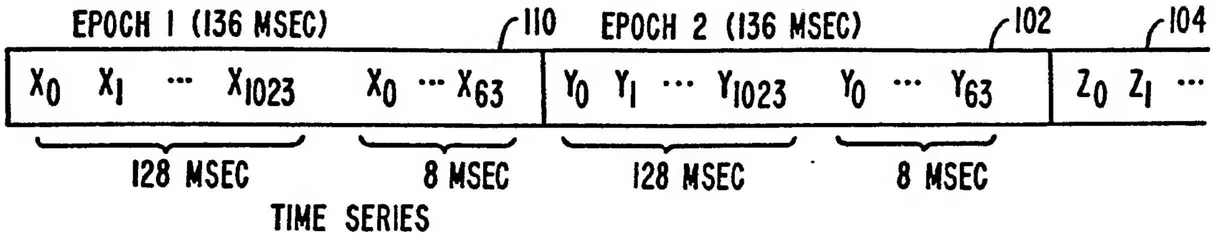


FIG._9.

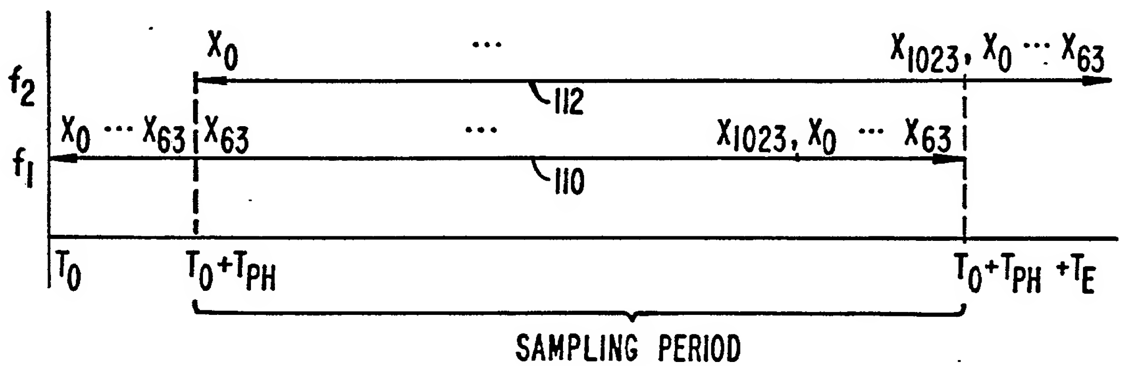


FIG._10.

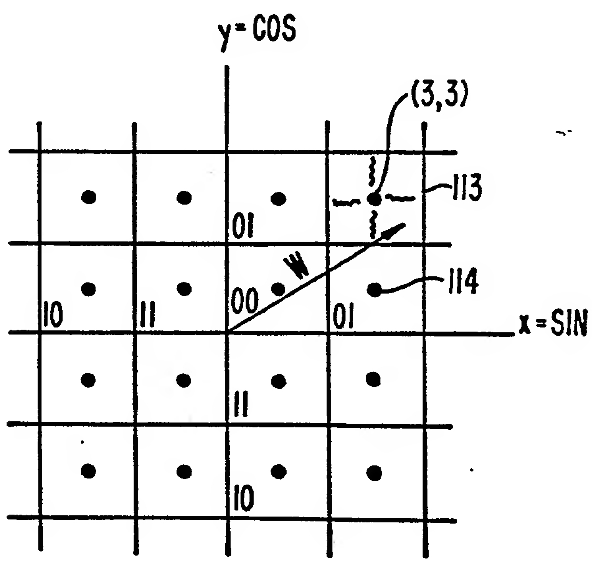


FIG._11.

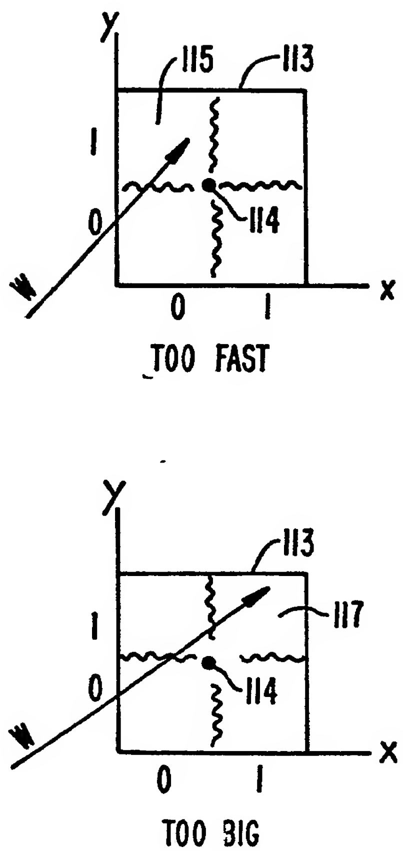

Y
115
113
114
116
117
118
1
0
X
W
TOO FAST
TOO SLOW
TOO BIG
TOO SMALL

FIG._12.

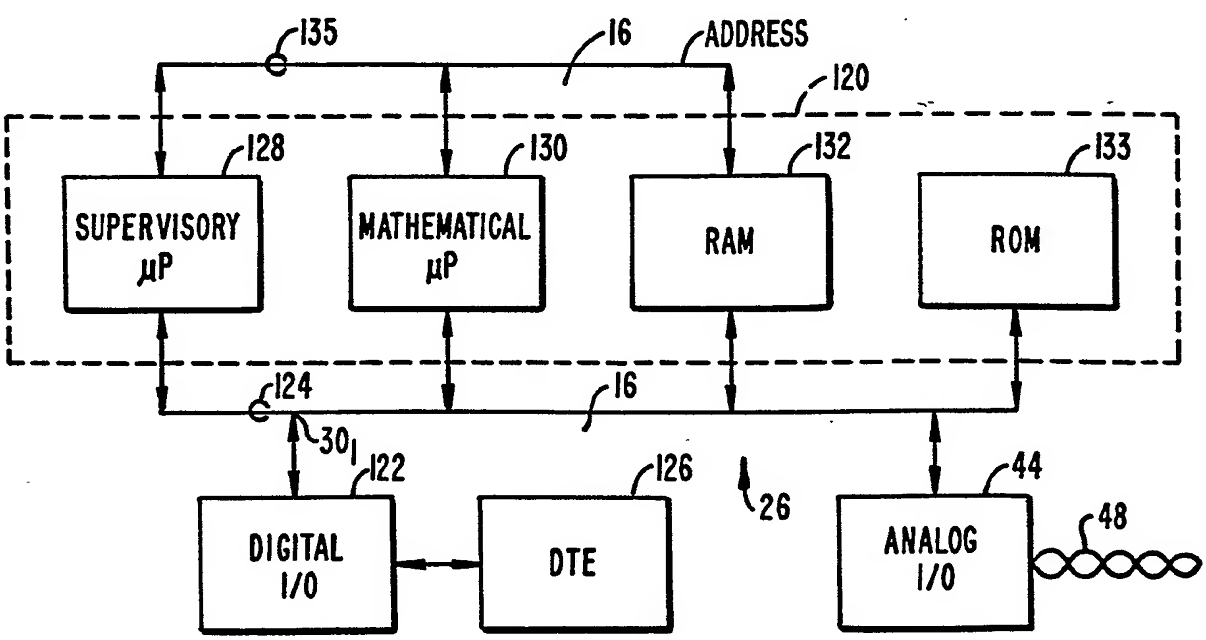

135
16
ADDRESS
120
128
130
132
133
SUPERVISORY µP
MATHEMATICAL µP
RAM
ROM
124
30
122
126
26
44
48
DIGITAL I/O
DTE
ANALOG I/O

FIG._13.

INTERNATIONAL SEARCH REPORT

International Application No PCT/US86/00983

I. CLASSIFICATION OF SUBJECT MATTER (if several classification symbols apply, indicate all) [3]

According to International Patent Classification (IPC) or to both National Classification and IPC

IPC (4): H04M 11/00;H04B 15/00,1/10;H04L 5/00,25/08;H04B 1/10

U.S. Cl.: 179/2DP; 375/39,58,99; 455/63

II. FIELDS SEARCHED

Minimum Documentation Searched [4]

Classification System	Classification Symbols
U.S.	179/2DP; 375/38,39,40,58,118; 370/16,108; 455/63,68+; 340/825.15

Documentation Searched other than Minimum Documentation to the Extent that such Documents are Included in the Fields Searched [5]

III. DOCUMENTS CONSIDERED TO BE RELEVANT [14]

Category *	Citation of Document, [16] with indication, where appropriate, of the relevant passages [17]	Relevant to Claim No. [18]
X,P	Telecommincations, Volume 19, No. 10, issued October 1985 (Dedham, Massachusetts), H.R. Johnson, "PC Communications: The Revolution Is Coming", see pages 58j to 58r.	1-17
A	US, A, 4,438,511 (Baran) 20 March 1984	1-17
A,P	US, A, 4,559,520 (Johnston) 17 December 1985	1-17
A	US, A, 4,206,320 (Keasler et al.) 03 June 1980	1-17
A	US, A, 3,810,019 (Miller) 07 May 1974	1-5,10-12,17
A	US, A, 4,328,581 (Harmon et al.) 04 May 1982	1-5,10-12,17
A	US, A, 3,971,996 (Motley et al.) 27 July 1976	6-8,13-15
A,P	US, A, 4,555,790 (Betts et al.) 26 November 1985	6-8,13-15

(cont'd)

* Special categories of cited documents: [15]

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"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

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"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

IV. CERTIFICATION

Date of the Actual Completion of the International Search [2]	Date of Mailing of this International Search Report [2]
17 June 1986	10 JUL 1986
International Searching Authority [1]	**Signature of Authorized Officer** [20]
ISA/US	Matthew E. Connors Matthew E. Connors

Form PCT/ISA/210 (second sheet) (October 1981)

International Application No. PCT/US86/00983

III. DOCUMENTS CONSIDERED TO BE RELEVANT (CONTINUED FROM THE SECOND SHEET)

Category *	Citation of Document, [16] with indication, where appropriate, of the relevant passages [17]	Relevant to Claim No [18]
A	US, A, 3,783,385 (Dunn et al.) 01 January 1974	1-5
A	US, A, 4,047,153 (Thirion) 06 September 1977	1-5
A	US, A, 4,494,238 (Groth, Jr.) 15 January 1985	1-5
A	US, A, 4,495,619 (Acampora) 22 January 1985	1-5,10-12,17
A	US, A, 4,484,336 (Catchpole et al.) 20 November 1984	1-5,10-12,17
A	US, A, 4,459,701 (Lamiral et al.) 10 July 1984	9,16,17
A	US, A, 3,755,736 (Kaneko et al.) 28 August 1973	9,16,17
A	US, A, 4,315,319 (White) 09 February 1982	1-5,10-12,17
A,P	US, A, 4,573,133 (White) 25 February 1986	1-5,10-12,17
A	US, A, 4,392,225 (Wortman) 05 July 1983	1-5,10-12,17

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International Bureau

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(30) Priority Data:
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(74) Agent: MILLER, Russell, B.; Qualcomm Incorporated, 6455 Lusk Boulevard, San Diego, CA 92121 (US).

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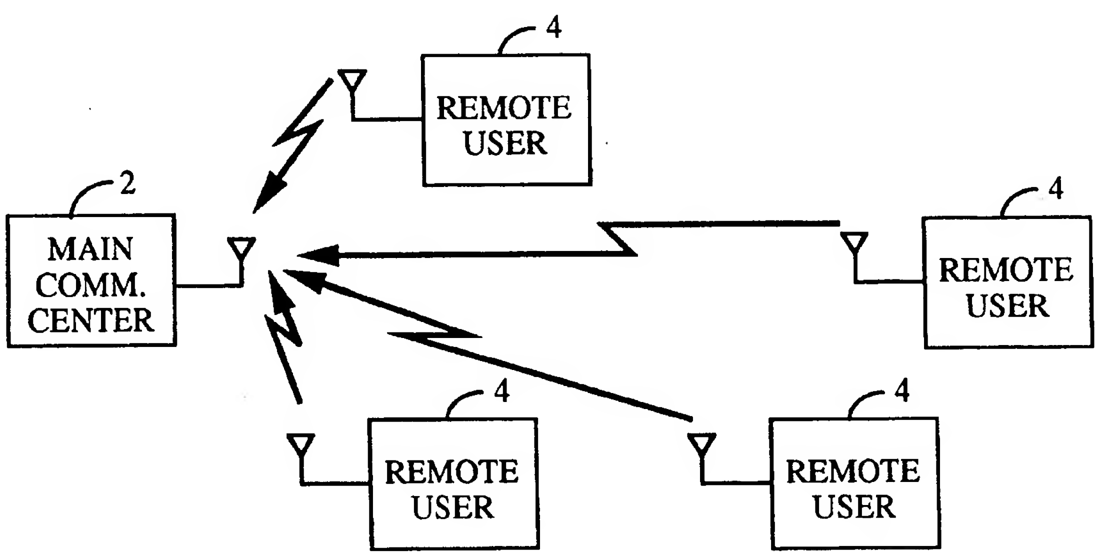
Published
With international search report.

(54) Title: METHOD AND APPARATUS FOR DETERMINING THE TRANSMISSION DATA RATE IN A MULTI-USER COMMUNICATION SYSTEM



(57) Abstract

A method and apparatus for controlling the data rates for communications to and from a base station (2) and a plurality of remote users (4). The usage of the communications resource whether the forward link resource, from base station (2) to remote users (4), or reverse link resource, from remote users (4) to base station (2), is measured. The measured usage value is compared against at least one predetermined threshold value and the data rates of communications or a subset of communications on said communications resource is modified in accordance with said comparisons.

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METHOD AND APPARATUS FOR DETERMINING THE TRANSMISSION DATA RATE IN A MULTI-USER COMMUNICATION SYSTEM

BACKGROUND OF THE INVENTION

I. Field of the Invention

The present invention relates to communications systems. More particularly, the present invention relates to a novel and improved method and apparatus for maximizing total average service quality to users in a multi-user communication system by controlling the data transmission rates to and from users of the multi-user communication system.

II. Description of the Related Art

The term "multiple access" refers to the sharing of a fixed communications resource by a plurality of users. A typical example of such a fixed communications resource is bandwidth. There are three basic ways to increase the throughput or data rate of an individual user accessing a communications resource. The first way is to increase the transmitters radiated power or alternatively to reduce system losses so that the received signal to noise ratio (SNR) is increased. The second way is to increase the allocation of bandwidth to the user. The third approach is to make allocation of the communications resource more efficient.

Some of the more common methods of providing multiple access to a communications resource involve both analog and digital communication modulation schemes. Such schemes include frequency division, time division and spread spectrum techniques. In frequency division multiple access (FDMA) techniques, each user is allocated one or more specific sub-bands of frequency. In time division multiple access (TDMA) techniques, periodically recurring time slots are identified, and for each segment of time each user is allocated one or more time slots. In some TDMA systems, users are provided a fixed assignment in time, and in other systems users may access the resource at random times. In spread spectrum communications, users share a common frequency band. Using frequency hopping (FH) modulation, the signal is modulated upon a carrier which changes in frequency according to a predetermined plan. In direct sequence (DS) modulation, the user signal is modulated with a pseudorandom code. In one type of code division multiple access (CDMA) technique which uses

direct sequence spread spectrum modulation, a set of orthogonal or nearly orthogonal spread spectrum codes (each using full channel bandwidth) are identified, and each user is allocated one or more specified codes.

In all multiple access schemes, a plurality of users shares a communications resource without creating unmanageable interference to each other in the detection process. The allowable limit of such interference is defined to be the maximum amount of interference such that the resulting transmission quality is still above a predetermined acceptable level. In digital transmission schemes, the quality is often measured by the bit error rate (BER) or frame error rate (FER). In digital speech communications systems, the overall speech quality is limited by data rate allowed for each user, and by the BER or FER.

Systems have been developed to minimize the data rate required for a speech signal while still providing an acceptable level of speech quality. If speech is transmitted by simply sampling and digitizing the analog speech signal, a data rate on the order of 64 kilobits per second (Kbps) is required to achieve a speech quality equivalent to that of a conventional analog telephone. However, through the use of speech analysis, followed by the appropriate coding, transmission, and resynthesis at the receiver, a significant reduction in the data rate can be achieved with a minimal decrease in quality.

Devices which employ techniques to compress speech by extracting parameters that relate to a model of human speech generation are typically called vocoders. Such devices are composed of an encoder, which analyzes the incoming speech to extract the relevant parameters, and a decoder, which resynthesizes the speech using the parameters which are received from the encoder over the transmission channel. As the speech changes, new model parameters are determined and transmitted over the communications channel. The speech is typically segmented into blocks of time, or analysis frames, during which the parameters are calculated. The parameters are then updated for each new frame.

A more preferred technique to accomplish data compression, so as to result in a reduction of information that needs to be sent, is to perform variable rate vocoding. An example of variable rate vocoding is detailed in U.S. Patent Application Serial No. 08/004,484 entitled "Variable Rate Vocoder," assigned to the assignee of the present invention and incorporated herein by reference. Since speech inherently contains periods of silence, i.e. pauses, the amount of data required to represent these periods can be reduced. Variable rate vocoding most effectively exploits this fact by

reducing the data rate for these periods of silence. A reduction in the data rate, as opposed to a complete halt in data transmission, for periods of silence overcomes the problems associated with voice activity gating while facilitating a reduction in transmitted information, thus reducing the overall interference in a multiple access communication system.

It is the objective of the present invention to modify the variability of the transmission rate of variable rate vocoders, and any other variable rate data source, in order to maximize utilization of the communications resource.

SUMMARY OF THE INVENTION

The present invention is a novel and improved method and apparatus for maximizing total average service quality to users in a multi-user communication system by controlling the data transmission rates to and from users of the multi-user communication system.

In the present invention, usage of the available communication resource is monitored. When the usage of the available communication resource becomes too great for a given communications link, and thus the quality falls below a predetermined limit, the data rate to or from the users is limited to free up a portion of the available communication resource. When the usage of the communications resource becomes small, the data rate to or from the users is allowed to increase above the previous limit.

For example, if the communications link from remote users to a main communications center, hereafter known as the reverse link, becomes overloaded, the main communications center transmits a signaling message requesting that the users, or selected ones of the users, decrease their average transmission data rate. At the remote user end, the signaling message is received and the transmission rate for the remote user is lowered in accordance with the signaling message.

The remote user, in the example, may be transmitting speech data or other digital data. If the user is transmitting speech data, then his transmission data rate may be adjusted using a variable rate vocoder as is described in above mentioned Application Serial No. 08/004,484. The present invention is equally applicable to any variable rate vocoding strategy when the remote user is transmitting speech data. If the user is transmitting digital data that is not speech data, the system can optionally instruct the remote user to modify the transmitted data rate for the specific digital data source.

On the communication link between the main communication center and the remote users, hereafter known as the forward link, the main communication center monitors the fraction of its total resource capacity that is being used for communicating to the remote users. If the fraction of the communications resource being used is too large, the main communication center will decrease the permitted average transmission data rate to each user or a subset of users. If the fraction of the communications resource being used is too small, the main communication center will permit the average data rate for each user to increase. As in the reverse link, the control of the data rate may be selective in nature based upon the nature of the data (speech or non-speech) being transmitted to the remote users.

BRIEF DESCRIPTION OF THE DRAWINGS

The features, objects, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters are identified correspondingly throughout and wherein:

Figure 1 is a block diagram illustrating multiple remote (mobile) users accessing a main communications center (cell base station);

Figure 2 is a block diagram illustrating the effects of a multi-cell (multiple main communications centers) environment on data reception at a remote (mobile) user;

Figure 3 is a graph of average service quality versus number of users at a particular average transmission data rate;

Figure 4 is a graph of average service quality versus number of users for three different average transmission data rates;

Figure 5 is a flowchart of the system monitor and control operation;

Figure 6 is a communication resource pie chart for forward link communications;

Figure 7 is a communication resource pie chart for reverse link communications;

Figure 8 is a communication resource pie chart illustrating the actions to be taken with respect to different fractions of resource usage;

Figure 9 is a communication resource pie chart illustrating conditions under which the data rate would be decreased by the control mechanism of the present invention

Figure 10 is a communication resource pie chart illustrating the effects of decreasing the data rate of the previous communications resource;

Figure 11 is a block diagram of the monitor and control system for controlling reverse link communications located at the main communications center;

Figure 12 is a block diagram of the monitor and control system for controlling reverse link communications located at the remote user; and

Figure 13 is a block diagram of the forward link monitor and control apparatus.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

Figure 1 illustrates the multi-user communications system communications between remotes users 4 and the main transmission center 2. In the exemplary embodiment these communications are conducted by means of a code division multiple access (CDMA) multi- user scheme, which is detailed in U.S. Patent Serial No. 4,901,307 entitled "Spread Spectrum Multiple Access Communication System Using Satellite of Terrestrial Repeaters (CDMA)," and U.S. Patent Serial No. 5,103,459 entitled "System and Method for Generating Signal Waveform in a CDMA Cellular Telephone System (CDMA)," both assigned to the assignee of the present invention and incorporated by reference herein. The communications that occur from the remote users to the main transmission center are referred to as reverse link communications. The communications link that enables communications from remote users 4 to a cell base station 2 is referred to as the reverse link. In a CDMA system, system user capacity is a function of the level of interference in the system.

Figure 2 illustrates the two main issues that result in the need for the control of the data rate to reduce interference and increase capacity. In the exemplary embodiment of a CDMA multi-cell cellular communications network, the main capacity limit on forward link communications is the interference from neighboring cells as illustrated by the propagation lines drawn from the cell base stations 12 and the single remote user or mobile station 10. The second effect on forward link capacity in the present embodiment is illustrated by the second propagation path 18 from a single cell base station to a mobile station 10. The cause of this effect, known as multipath, is reflection off of obstruction 16 which may take the form of a

building, a mountain, or any other object that is capable of reflecting electromagnetic waves.

In the exemplary embodiment, interference is received by remote user 10 from cell base stations 12 which are not communicating with the remote user, and interference is received by multipath signals from obstruction 16. In the exemplary embodiment, the operation of a group of cells is overseen by the system controller 14 that provides the data to and from a public telephone switching network (not shown). These communications are referred to as forward link communications.

In systems like time division multiple access (TDMA) and frequency division multiple access (FDMA), a "hard" capacity limit exists due to the finite number of time slot or frequency sub-band divisions, respectively. When all of the time slots or subbands are allocated to users, the "hard" capacity limit is reached and service to any additional user is impossible. Though the users that have accessed the system before the capacity limit remain unaffected by any excluded users, the average quality of service to all users drops beyond the capacity limit since the quality of service for each additional user denied service is zero.

In multiple access schemes such as code division multiple access (CDMA) and random access systems like ALOHA and slotted ALOHA systems, a "soft" capacity limit exists. For these types of multiple access systems, the increase of the number of system users beyond a capacity limit causes a decrease in the quality of service to all users of the system. In a CDMA system, the transmissions of each user are seen as interference, or noise, to each other user. Beyond the soft capacity limit of a CDMA system, the noise floor becomes large enough to cause the predetermined allowable BER or FER to be exceeded. In random access schemes, each additional user increases the probability of a message collision. Beyond a capacity limit the message collisions grow so frequent that the need for retransmission or the resultant lost data causes the communication quality of all users to suffer.

Figure 3 is a graph of the average quality of service to users of such a multiple access communication system versus the number of users occupying the system, given a specified average data rate for all users. The average quality (Q_{ave}) of service is defined as:

$$Q_{ave} = \frac{1}{N}\sum_{i=1}^{N} Q_i \tag{1}$$

where Q_i is quality of service to user i and N is the number of users on the system.

Figure 3 also illustrates a quality line above which the average service quality is acceptable and below which the service quality is unacceptable. The intersection of the quality line with the plot of quality versus number of users curve defines the capacity limit of the system at the data rate of the system. In the exemplary embodiment of a CDMA system, messages are transmitted in 20 ms frames, and a tolerable frame error rate of 1% dictates the position of the quality line in the exemplary embodiment. It is understood that different frame sizes and error rates are equally applicable to the present invention.

Figure 4 illustrates three plots 20, 22, and 24 of average quality of service versus number users for three progressively decreasing average data rates. Plot 20 corresponds to the quality curve for a high average data rate, plot 22 corresponds to the quality curve for a moderate average data rate, and plot 24 corresponds to the quality curve for a low average data rate.

The first important feature in the plots is that the intersection of the plots with the vertical axis is progressively lower for lower link data rates. Below capacity limits, higher allowable data rates correspond to higher quality, since a high data rate allows more precise quantization of the parameters in the variable rate speech coder, resulting in cleaner sounding speech.

The second important feature in the plots is the intersections of the quality line with the three plots. The intersections of the quality line with each of the curves 20, 22 and 24 provides the capacity limit for the system at the respective data rates of curves 20, 22 and 24. The system capacities labeled CAP A, CAP B, and CAP C are the number of users that can access the system at the data rates of each of curves 20, 22 and 24. The capacity limit at a given data rate is obtained by dropping a vertical line, as shown in the diagram, from the intersection of the plot and the quality line to the horizontal axis representing the number of users. The capacity of the system increases for a fixed quality level as the data rate decreases.

Figure 5 is a flowchart illustrating the method of maximizing the average quality by controlling the data rate of transmission on the system. At block 30 the amount of communications resource that is in use is determined, based on the number of users accessing the system on the given link and the data rate transmitted by each user. The usage value computed in block 30 is passed to block 32. In block 32 the usage value is compared against a lower threshold. If the usage value is below the lower threshold

then the operation goes to block 34 where it is determined if the link is operating at a predetermined data rate maximum. If the system is operating at the predetermined data rate maximum, the operation moves to block 38 and no action is taken. If the system is operating below the predetermined data rate maximum, the operation proceeds to block 36 and the link data rate is increased.

If back at block 32 it is determined that the link usage is not too low, the operation proceeds to block 40 where the usage is compared against an upper threshold. If in block 40 the link usage is determined to be below the upper threshold, the operation proceeds to block 41 and no action is taken. If on the other hand, the link usage exceeds the upper threshold in block 40, the operation proceeds to block 42. In block 42, the system data rate is compared against a predetermined minimum. If the system data rate is greater than this predetermined minimum then the operation proceeds to block 44 where the link data rate is decreased.

If at block 42 the link data rate was determined to be equal to the minimum link data rate then the operation proceeds to block 46. At block 46 the system compares the usage to a predetermined usage maximum. If the communications resource is exhausted, that is the usage is equal to the predetermined maximum, then the operation proceeds to block 48 and access by any additional users is blocked. If the usage is below the predetermined usage maximum then, then operation proceeds to block 50 and no action is taken.

In TDMA systems, data rates can be modified by spreading data of a given user among a plurality of allocated time slots or combining the data of a plurality of users with selected ones of allocated time slots. In an alternative implementation variable data rates could be achieved in a TDMA system by allocating time slots of varying length to different users. Similarly, in FDMA systems data rates can be modified by spreading data of a given user among a plurality of allocated frequency sub-bands or combining the data of a plurality of users with selected ones of allocated frequency sub-bands. In an alternative implementation variable data rates in a FDMA system could be achieved by allocating varying frequency sub-bands sizes to different users.

In random access systems the probability of message collisions is proportional to the amount of information each user needs to send. Therefore, the data rate can be adjusted directly by sending varying size packets of data or by sending the packets at varying time intervals between transmission.

In the exemplary embodiment using a CDMA system, the amount of data necessary for transmission of speech is adjusted by use of a variable rate vocoder as detailed in Application Serial No. 08/004,484 mentioned above. The variable rate vocoder of the exemplary embodiment, provides data at full rate, half rate, quarter rate and eighth rate corresponding to 8Kbps, 4Kbps, 2Kbps and 1Kbps, but essentially any maximum average data rate can be attained by combining data rates. For example, a maximum average rate of 7Kbps can be attained by forcing the vocoder to go to half rate every fourth consecutive full rate frame. In the exemplary embodiment, the varying size speech data packet, is segmented and segments are provided at randomized times as is detailed in U.S. Patent Application Serial No. 07/846,312 entitled "Data Burst Randomizer," assigned to the assignee of the present invention and incorporated by reference herein.

A useful way of looking at the issue of communications resource capacity is to view the available communications resource as a pie chart, where the whole pie represents the complete exhaustion of the communication resource. In this representation sectors of the pie chart represent fractions of the resource allocated to users, system overhead, and unused resource.

In a TDMA or FDMA system the whole of the pie chart may represent the number of available time slots or frequency sub-bands in a given allocation strategy. In a random access system, the whole of the pie chart may represent the message rate that is acceptable before message collisions grow so great as to make the transmission link unacceptable. In the exemplary embodiment of a CDMA system, the whole of the pie chart represents the maximum tolerable noise floor wherein the overhead and signal from all other users appear as noise in the reception of the message data to and from the remoter users. In any system configuration, referring back to Figure 3, the whole of the resource pie represents the intersection of the quality line with the average quality versus number of users plot.

Figure 6 illustrates an example of a general forward link capacity pie chart. The first sector of the resource pie labeled OVERHEAD represents the portion of the transmission signal that does not carry message information. The OVERHEAD fraction of the pie represents the transmission of non-message non-user-specific data and in the exemplary embodiment is a fixed fraction of the communication resource though in other systems this overhead may vary with the number of users or other factors. The OVERHEAD may include base station identification information, timing information and base station setup information among other things. The

OVERHEAD may include pilot channel usage of the communications resource. An example of a pilot channel is detailed in U.S. Patent Serial No. 5,103,459, entitled "System and Method for Generating Signal Waveforms in a CDMA Cellular Telephone System (CDMA)," assigned to the assignee of the present invention and incorporated herein by reference. Each of the following sectors numbered 1-20 represents a the message information directed to a particular user, where the users are numbered 1-20. The last sector of the pie, moving in a clockwise direction, is labeled with a B. The sector labeled with a B represents the remaining fraction of available communication resource before unacceptable link degradation occurs.

Figure 7 is a resource pie chart for the reverse link communications. This pie chart represents the information received at the main transmission center or base station from the remote users. The only significant difference between this pie chart and the previous pie chart is in the reverse link there is no fixed OVERHEAD resource. It should also be noted that in the preferred embodiment each user uses the same fraction of communication resource in order to maximize the quality of service to all users. The method and apparatus for maintaining the condition wherein all users use the same fraction of received communication resource is detailed in U.S. Patent No. 5,056,109 entitled "Method and Apparatus for Controlling Transmission Power in a CDMA Cellular Telephone System" assigned to the assignee of the present invention and incorporated by reference herein. In this approach, each remote user transmits at a power level such that it is received at the base station as all other remote users. Preferably, each remote user transmits at a minimum power level necessary to insure a quality communication link with a base station.

Figure 8 is an action pie chart that represents the actions to be followed with respect to the resource pie charts. Labeled on the pie chart of Figure 7 are three points, a point marked INCREASE RATE, a point marked DECREASE RATE and a point marked BLOCK ADDITIONAL USERS. If the fraction of the resource pie for a given link exceeds the point marked DECREASE RATE, the transmission rate on that link should be decreased to improve the quality of service to the users. For example, if the data rate corresponding to plot 20 in figure 4 was being transmitted by all users and the number of users became greater than CAP A, the data rate would be decreased, and the system would then operate on plot 22 in figure 4. If the fraction of the resource pie for a given link falls below the point marked INCREASE RATE, the transmission rate on that link should be increased to

improve the quality of service to the users. For example, if the data rate corresponding to plot 22 in figure 4 was being transmitted by all users and the number of users dropped below CAP A, the data rate would be increased and the system would operate on plot 20 in figure 4. If the pie reaches the point marked BLOCK ADDITIONAL USERS then any additional users should be blocked from accessing the system. Note that the only way the system would reach the BLOCK ADDITIONAL USERS point is by going through the DECREASE RATE point which implies that the rate could not be further decreased.

Figures 9 and 10 illustrate the effects of decreasing the transmission rate on the resource allocation. In Figure 8, the addition of user 20 has caused the resource allocation to surpass the point at which the transmission rate should be decreased. At this point the transmission rate is decreased and the resource pie for the same users looks like Figure 9. Notice the unused portion of the resource pie labeled B is large enough to permit additional users to access the communication resource. Thus, additional users can access the communication system until the system requires the transmission rate to be decreased again. This process will continue until the rate is at a minimum. If this occurs, the system allows the pie to fill entirely and then any new users are prevented from accessing the system.

In contrast as users leave the communication resource then the fraction of the communication resource that is used decreases below the INCREASE RATE point and the system will increase the transmission rate. This can continue until the transmission rate is increased to a maximum rate or until no users are accessing the communication resource.

Figure 11 illustrates a block diagram for the monitor and control of the reverse link communication resource usage at the main communications center, which may include the cell base station and the system controller. The signals from the remote users are received at receive antenna 60. The received signals are provided to receiver 62 which provides the data in analog or digital form to energy computation element 66 and demodulators 64. The computed energy value from energy computation element 66 is provided to rate control logic 68 which compares the received signal energy to a series of thresholds. In response to the comparisons, rate control logic 68 provides a rate control signal to microprocessors 70 when the signal energy is above an upper threshold or is below a lower threshold. In other embodiments, the rate control logic 68 could also be responsive to external factors which may affect the

performance of the communications channel, such as weather conditions, etc.

The received signal from receiver 62 are provided to demodulators 64, where it is demodulated and the data for a specific user is extracted and provided to the corresponding microprocessor 70. In the exemplary embodiment, as detailed in U.S. Patent Application Serial No. 07/433,031 entitled "Method and System for Providing a Soft Handoff in Communication in a CDMA Cellular Telephone System" assigned to the assignee of the present invention and incorporated by reference herein, the received data is provided by microprocessors 70 to selector cards (not shown) in a system controller 14 that selects a best received data from received data from a plurality of main communication centers (cells), each of which contains a receiver 62 and a demodulator 64, and decodes the best received data using a vocoder (not shown). The reconstructed speech is then provided to a public telephone switching network (not shown).

In addition, microprocessors 70 receive data for forward link transmission from the vocoders (not shown) through the data interface. Microprocessors 70 combine the reverse link rate control signal, when present, with the outgoing forward link data to provide composite data packets to modulators 72. In a preferred embodiment, ones of microprocessors 70 would selectively combine the reverse link rate control when present to with outgoing forward link data. In the preferred embodiment, ones microprocessors 70 are responsive to a signal indicative of overriding conditions where upon the reverse rate control signal is not combined with the outgoing forward link data. In alternate embodiment, certain ones of said microprocessors 70 would not be responsive to the reverse link rate control signal. Modulators 72 modulate the data packets and provide the modulated signals to summer 74. Summer 74 sums the modulated data and provides it to transmitter 76 where amplified and provided to transmission antenna 78.

Figure 12 illustrates a block diagram of the remote user apparatus of the present invention for responding to the rate control signal provided in the exemplary embodiment by main transmission center 2 in figure 1. On the receive path, the signal that comprises encoded speech data and/or signaling data is received at antenna 90, which also serves as the transmission antenna by means of duplexer 92. The received signal is passed through duplexer 92 to demodulator 96. The signal is then demodulated and provided to microprocessor 98. Microprocessor 98 then decodes the signal and passes the speech data and any rate control data that

is sent by the base station to the variable rate vocoder 100. Variable rate Vocoder 100 then decodes the encoded packet of speech data provided from microprocessor 98 and provides the decoded speech data to codec 102. Codec 102 converts the digital speech signal into analog form and provides the analog signal to speaker 106 for playback.

On the transmit path of the remote user, a speech signal is provided through microphone 106 to codec 102. Codec 102 provides a digital representation of the speech signal to the variable rate vocoder 100 which encodes the speech signal at a rate determined in the exemplary embodiment in accordance with the speech activity and the received rate signal. This encoded speech data is then provided to microprocessor 98.

In the exemplary embodiment, the rate control signal is a binary signal indicating to the remote user to increase or decrease the maximum data rate. This adjustment of the data rate is done in discrete levels. In the exemplary embodiment, the remote user will increase or decrease its maximum transmission rate by 1000 bps upon receipt rate control signaling from the cell base station. In practice, this reduces the overall average data rate by 400 to 500 bps, since the vocoder is only encoding the speech at the maximum rate 40-50% of the time in a normal two-way conversation. In the exemplary embodiment, the silence between words is always encoded at the lower data rates.

For example, if the remote user is currently operating with a maximum transmission data rate of full rate or rate 1 (8 Kbps), and a signal decrease its maximum data rate is received, the maximum transmission data rate will be decreased to 7/8 (7 Kbps) by forcing every fourth consecutive full rate frame of data to be encoded at half rate (4Kbps). If on the other hand, the remote user is operating under control of the cell base station at a maximum transmission rate of 3/4 (6 Kbps) and the cell base station signals the remote user to increase its maximum data rate, then the remote user will use a rate 7/8 (7 Kbps) as a maximum transmission data rate. In a simplified embodiment the rates could simply be limited to one of the discrete rates provided by variable rate vocoder 100 (i.e.. rates 1, 1/2, 1/4 and 1/8).

Microprocessor 98, also, receives non-speech data that can include signaling data or secondary data such as facsimile, modem, or other digital data that needs communication to the cell base station. If the digital data being transmitted by the remote user is of a form not conducive to variable rate transmission (i.e. some facsimile or modem data) then microprocessor

98 can decide based upon the service option of the remote user whether to vary the transmission rate in response to the rate control signal.

Modulator 108 modulates the data signal and provides the modulated signal to transmitter 110 where it is amplified and provide through duplexer 92 to antenna 90 and transmitted over the air to the base station. It is also envisioned in the present invention that the remote user could monitor the reverse link communication resource and respond in an open loop manner to adjust its transmission rate.

Figure 13 illustrates a block diagram of an exemplary forward link rate control apparatus. Speech data is provided to vocoders 120 where the speech data is encoded at a variable rate. In the present invention the encoding rate for the speech data is determined in accordance with the speech activity and a rate control signal when present. The encoded speech is then provided to microprocessors 122, which also may receive non-speech data from an external source (not shown). This non-speech data can include signaling data or secondary data (facsimile, modem or other digital data for transmission). Microprocessors 122 then provide data packets to modulators 124 where the data packets are modulated and provided to summer 126. Summer 126 sums the modulated signal from modulators 124 and provides the sum signal to transmitter 128 where the signal is mixed with a carrier signal, amplified and provided to antenna 130 for transmission.

The summed modulated signal from summer 126 is also provided to energy computation unit 132. Energy computation unit 132 computes the energy of the signal from summer 126 for a fixed time period and provides this energy estimate to rate control logic 134. Rate control logic 134 compares the energy estimate to a series of thresholds, and provides a rate control signal in accordance with these comparisons. The rate control signal is provided to microprocessors 122. Microprocessors 122 provide the rate control signal to vocoders 120 for control of the maximum data rate of speech data. Optionally, microprocessors 122 can also use the rate control signal to control the data rate of non-speech data sources (not shown). the rate control signal can be provided selectively to ones of microprocessors 122 or alternately selects ones of microprocessors 122 can be responsive to a globally provided rate control signal.

The open loop form of control on the forward link described above can also operate in a closed loop, which can be responsive to signals from the remote stations indicative of capacity limits being reached, such as high frame error rates or other measurable quantities. Rate control logic 134 can

be responsive to external interferences of various kinds which may also affect the performance of the communications channel.

The previous description of the preferred embodiments is provided to enable any person skilled in the art to make or use the present invention. The various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without the use of the inventive faculty. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WE CLAIM:

CLAIMS

1. In a communications network wherein a plurality of remote users each having a transmitter communicate message signals to a communications center having a receiver, a subsystem for optimizing communications quality in accordance with system usage and capacity comprising:

monitor means for determining said system usage and conditionally providing a rate control signal in accordance with said usage level; and

a plurality of response means each collocated with a corresponding one of said remote users for receiving said rate control signal and adjusting said data rate of said corresponding one of said remote users in accordance with.

2. The subsystem of Claim 1 wherein said monitor means is collocated with said transmission center, said subsystem further comprising:

communications center transmitter means for transmitting messages to said remote users and for transmitting said rate control signal to said remote users; and

plurality of remote receiver means, each of said plurality of remote receiver collocated with a corresponding one of said remote users for receiving said rate control signal and for providing said rate control signal to a corresponding one of said response means.

3. The subsystem of Claim 1 wherein said monitor means determines said system usage by measuring the energy of said message signals for a predetermined time period.

4. The subsystem of Claim 1 wherein said response means comprises:

processor means for receiving said rate control signal and providing a rate command signals in response to said rate control signal; and

variable rate vocoder means for receiving speech data and said rate command signals and encoding said speech data at a rate in accordance with said command signals.

5. The subsystem of Claim 4 wherein said variable rate vocoder means further encodes said speech data in accordance with the energy of said speech data.

6. The subsystem of Claim 4 wherein said processor means is further for receiving non-speech data for transmission and for providing said non-speech data at a rate in accordance with said rate control signal.

7. A variable rate transceiver comprising:

a receiver for receiving a signal comprising message data and a rate control command;

a variable rate vocoder for receiving speech data and encoding said speech data in accordance with said rate control command; and

a transmitter for transmitting said encoded speech data.

8. The variable rate transceiver of Claim 7 further comprising:

a demodulator disposed between said receiver and said variable rate vocoder for demodulating said received signal; and

a processor disposed between demodulator and said variable rate vocoder for receiving said demodulated signal and separately providing said message data and said rate control command.

9. The variable rate transceiver of Claim 8 wherein said processor is further for receiving non-speech data for transmission.

10. The variable rate transceiver of Claim 7 further comprising a modulator disposed between said variable rate vocoder and said transmitter for modulating said encoded speech data.

11. The variable rate transceiver of Claim 7 further comprising a modulator disposed between said variable rate vocoder and said transmitter for modulating said encoded speech data.

12. At a base station, an apparatus for controlling the user capacity of said base station comprising:

usage determination means for measuring usage of said base station;

rate control means for comparing said measured usage against at least one predetermined value and selectively providing a rate control signal in accordance with said comparisons; and

transmitter means for transmitting said rate control signal.

13. The apparatus of Claim 12 further comprising processor means for receiving message data for transmission to said remote users and said rate control signal and combining said message data with said rate control signal to provide a composite data packet.

14. The apparatus of Claim 13 further comprising a modulator means disposed between said processor means and transmitter for modulating said composite data packet.

15. In a communication system wherein a base station communicates messages on a forward link with a plurality of remote users an apparatus of controlling the data rate of said message communications, comprising:

usage determination means for determining a usage value of said forward link;

rate control logic means for receiving said usage value, comparing said usage value to at least one predetermined threshold value and conditionally providing a rate control signal in accordance with said comparisons; and

at least one variable rate data source for providing data at a rate in accordance with said rate control signal.

16. The apparatus of Claim 15 wherein said at least one variable rate data source comprises at least one variable rate vocoder means for encoding speech data at variable rates.

17. The apparatus of Claim 15 wherein said usage determination means measures the energy of a signal for transmission to said remote users.

18. A method for optimizing usage of a communications resource, comprising the steps of:

measuring said usage of said communications resource;

comparing said measured usage against at least one predetermined threshold; and

adjusting data rates of communications on said communications resource in accordance with said comparisons.

19. The method of Claim 18 wherein said step of comparing said measured usage against at least one predetermined threshold comprises comparing said usage against a predetermined high usage threshold, and wherein said step of adjusting data rates of communications on said communications resource comprises decreasing the data rate of communications when said usage exceeds said high usage threshold.

20. The method of Claim 18 wherein said step of comparing said measured usage against at least one predetermined threshold comprises comparing said usage against a predetermined low usage threshold, and wherein said step of adjusting data rates of communications on said communications resource comprises increasing the data rate of communications when said usage falls below said low usage threshold.

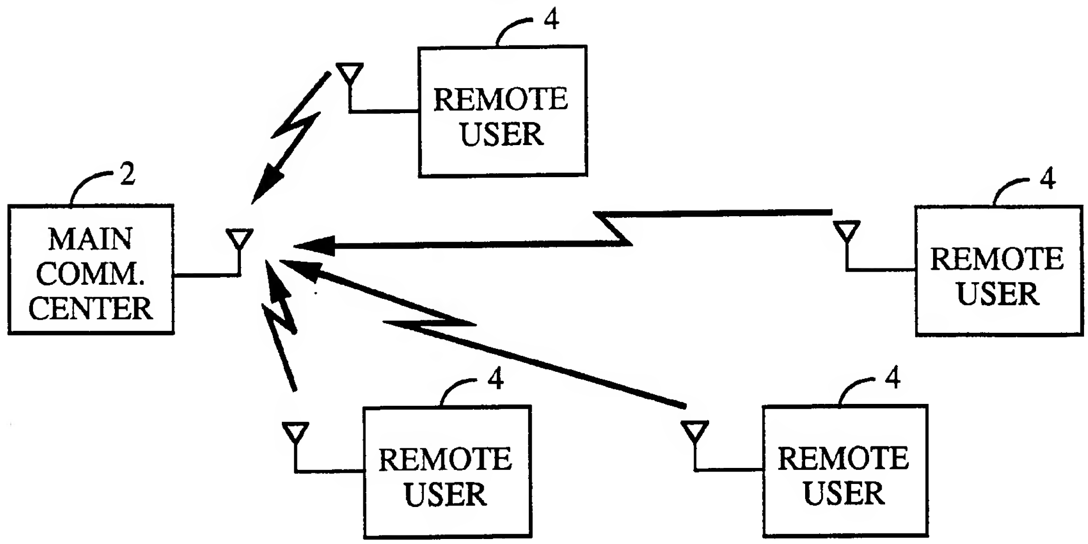

4
REMOTE USER
2
MAIN COMM. CENTER
4
REMOTE USER
4
REMOTE USER
4
REMOTE USER

FIG. 1

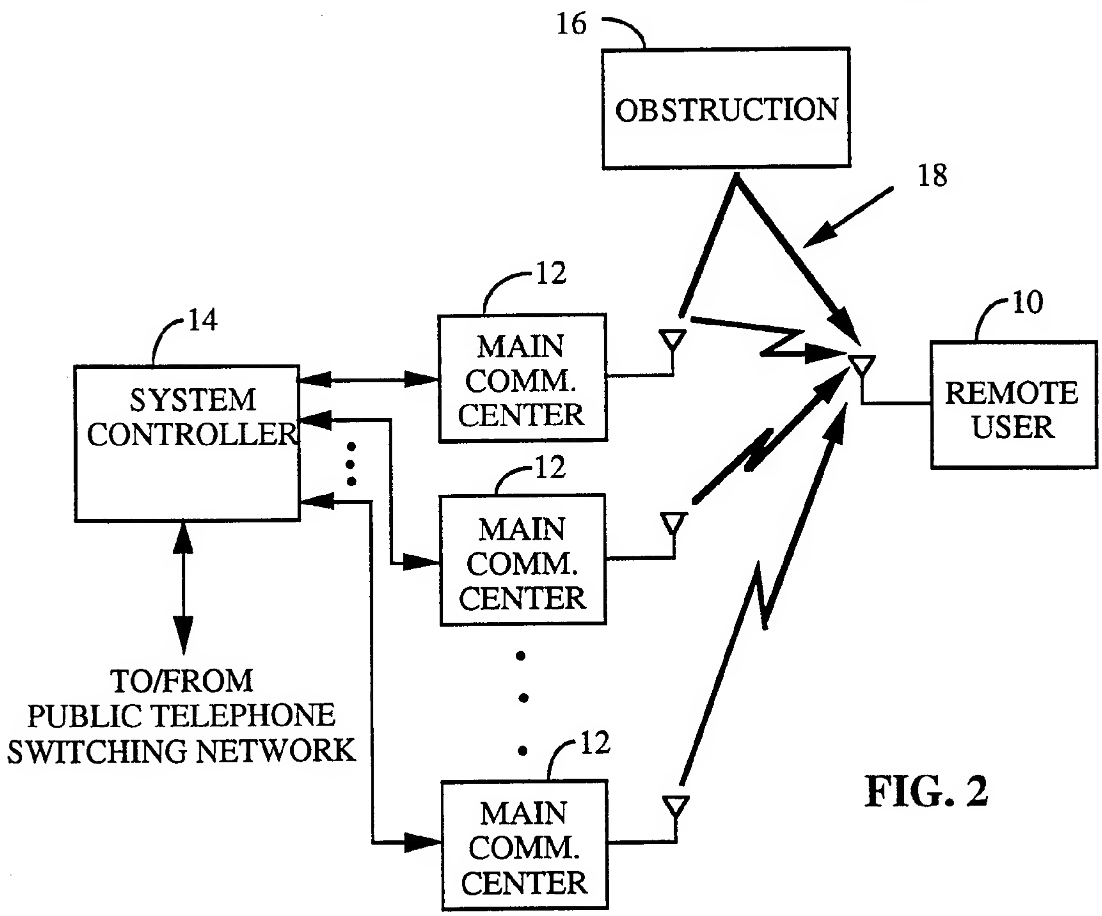
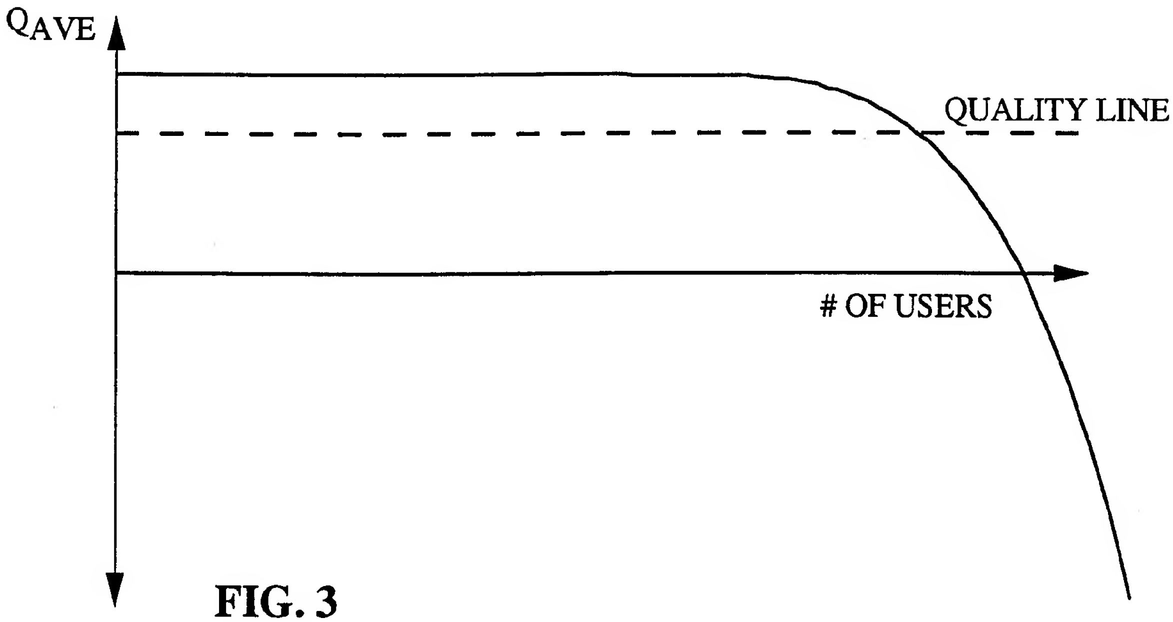
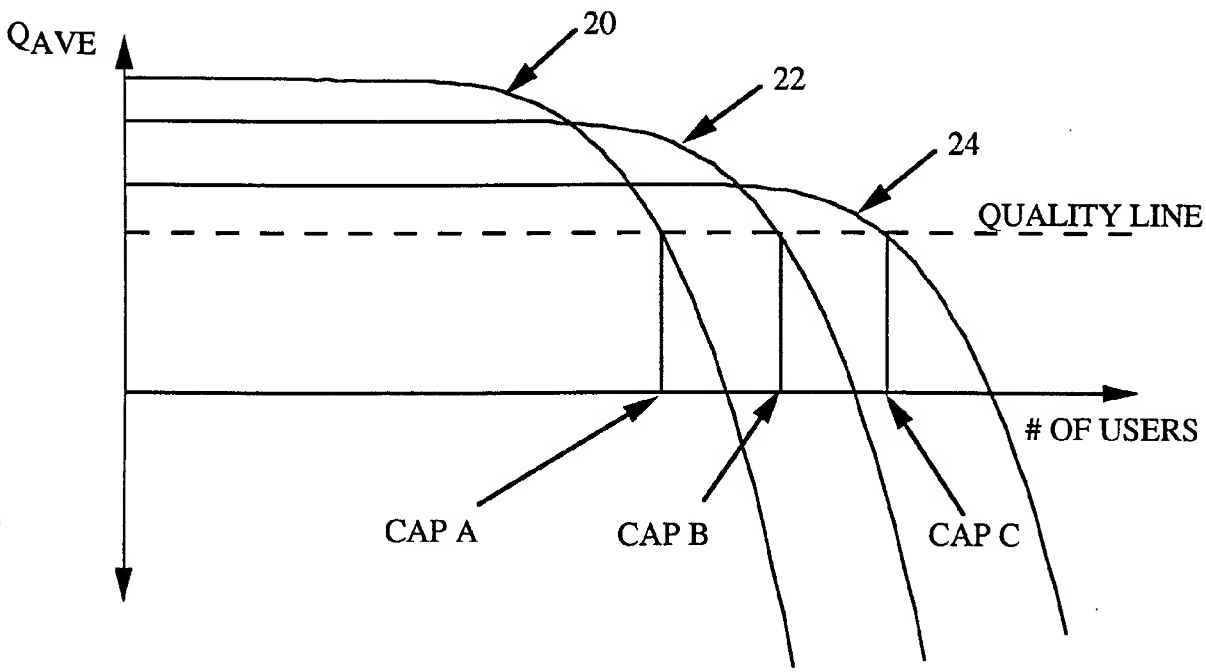

16
OBSTRUCTION
18
12
MAIN COMM. CENTER
14
SYSTEM CONTROLLER
10
REMOTE USER
12
MAIN COMM. CENTER
TO/FROM PUBLIC TELEPHONE SWITCHING NETWORK
12
MAIN COMM. CENTER

FIG. 2



QAVE
QUALITY LINE
OF USERS

FIG. 3



QAVE
20
22
24
QUALITY LINE
OF USERS
CAP A
CAP B
CAP C

FIG. 4

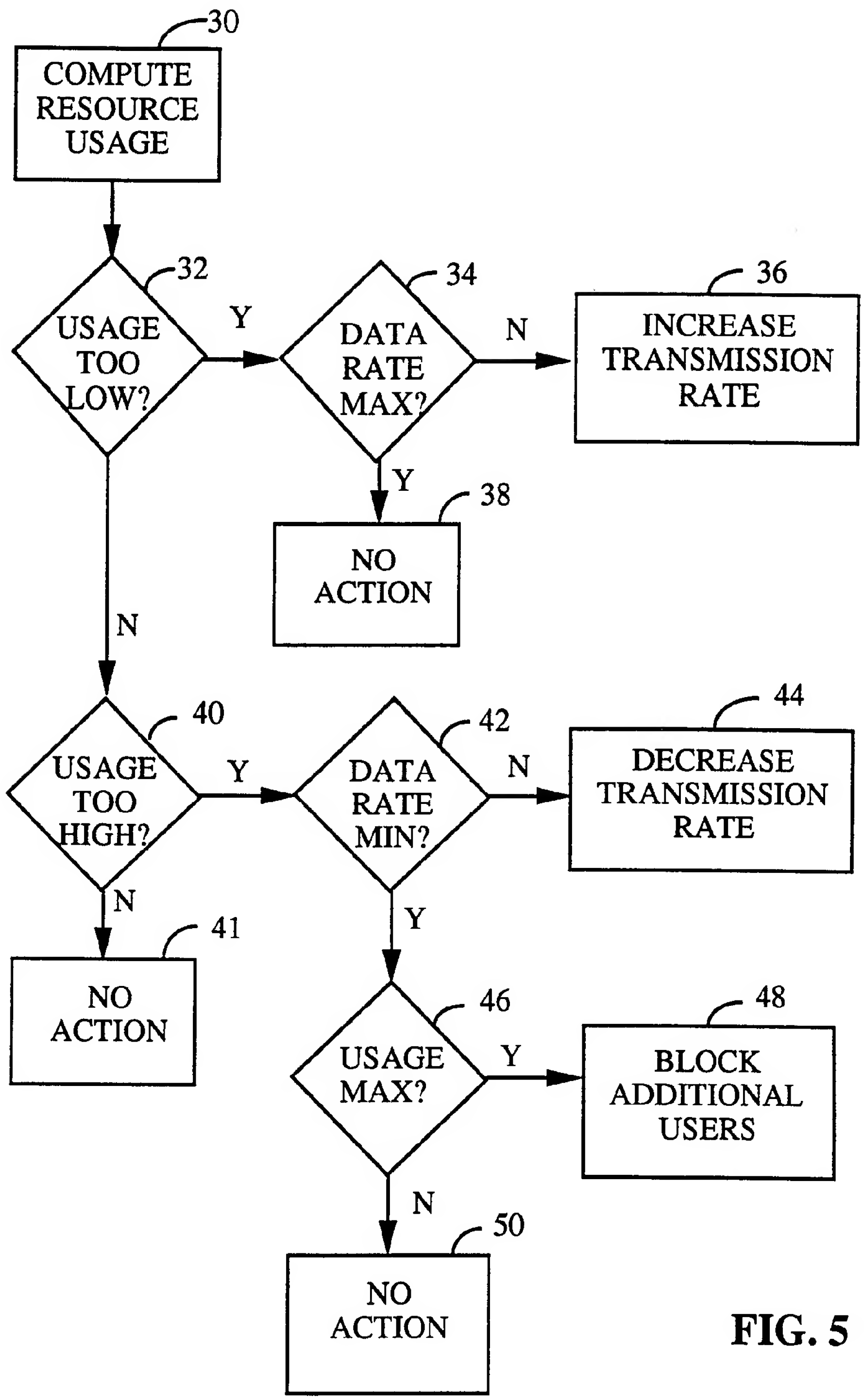

30
COMPUTE RESOURCE USAGE
32
USAGE TOO LOW?
Y
34
DATA RATE MAX?
N
36
INCREASE TRANSMISSION RATE
Y
38
NO ACTION
N
40
USAGE TOO HIGH?
Y
42
DATA RATE MIN?
N
44
DECREASE TRANSMISSION RATE
N
41
NO ACTION
Y
46
USAGE MAX?
Y
48
BLOCK ADDITIONAL USERS
N
50
NO ACTION

FIG. 5

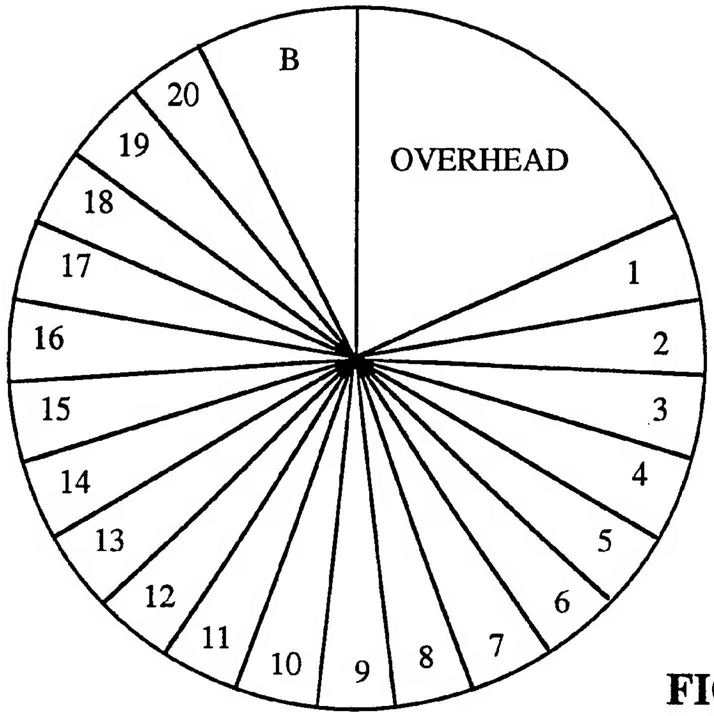
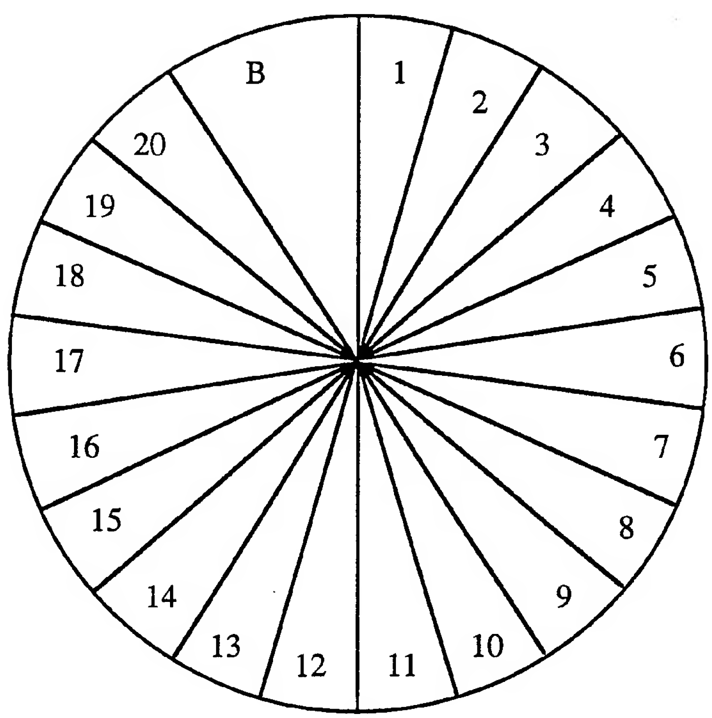

B
OVERHEAD
20
19
18
17
16
15
14
13
12
11
10
9
8
7
6
5
4
3
2
1

FIG. 6


B
1
2
3
4
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6
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8
9
10
11
12
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16
17
18
19
20

FIG. 7

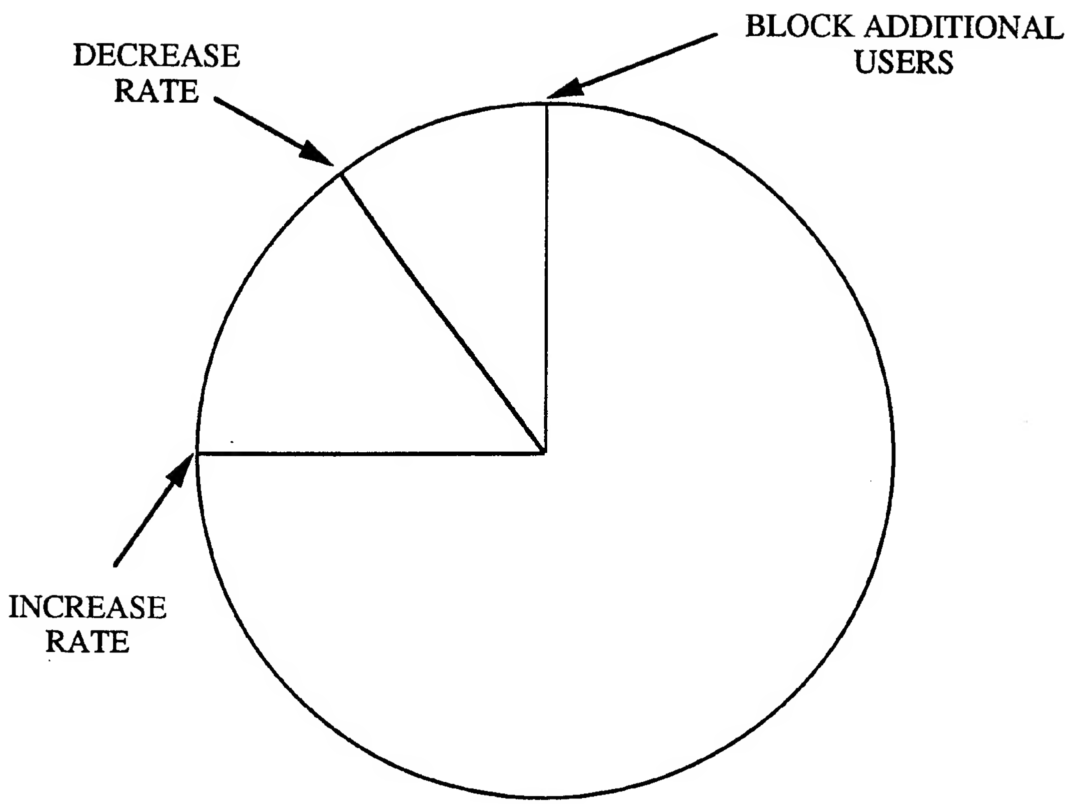


FIG. 8

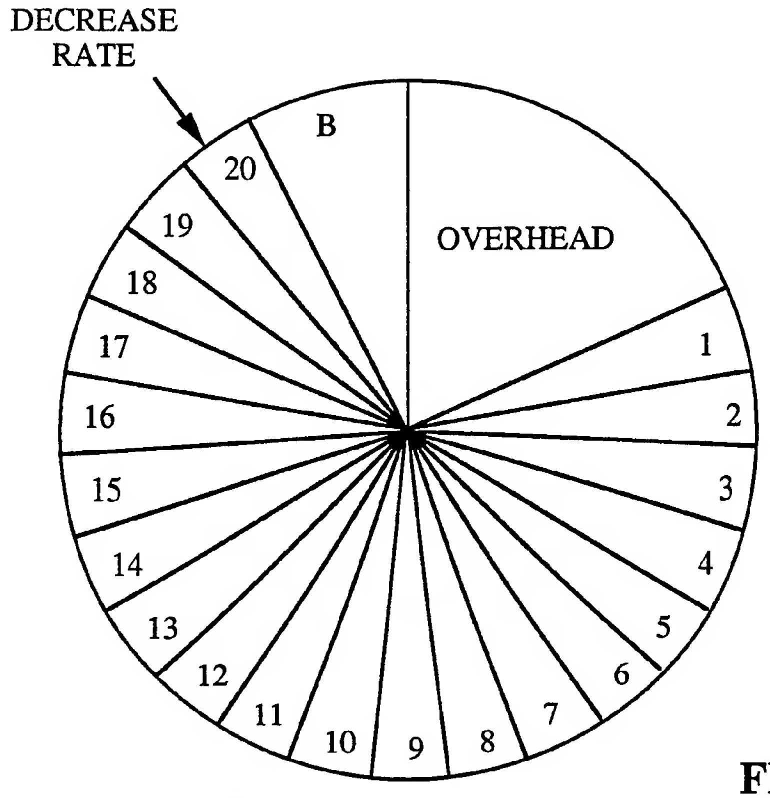
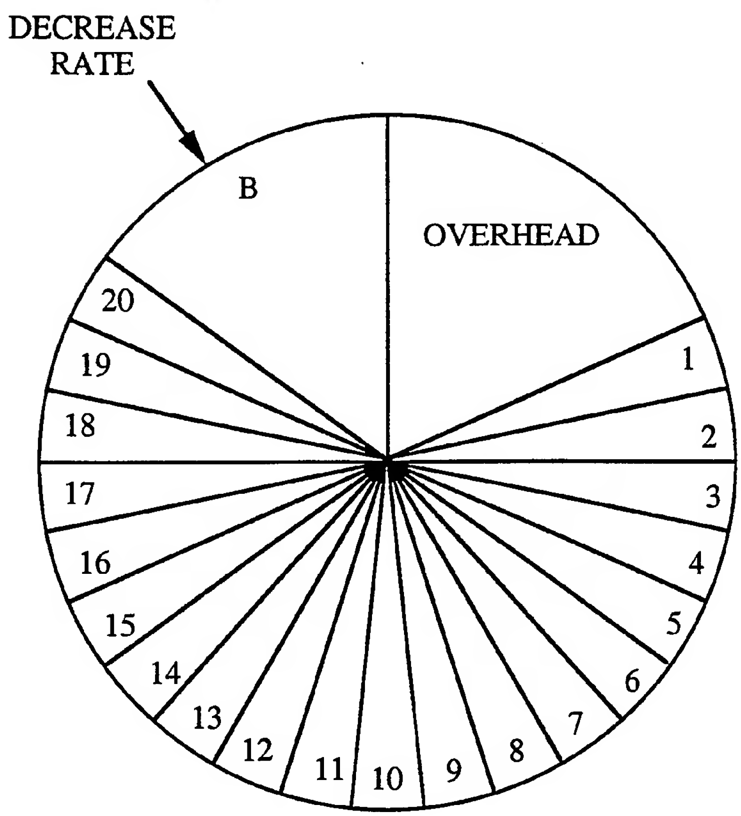

DECREASE
RATE
B
OVERHEAD
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16
17
18
19
20

FIG. 9


DECREASE
RATE
B
OVERHEAD
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16
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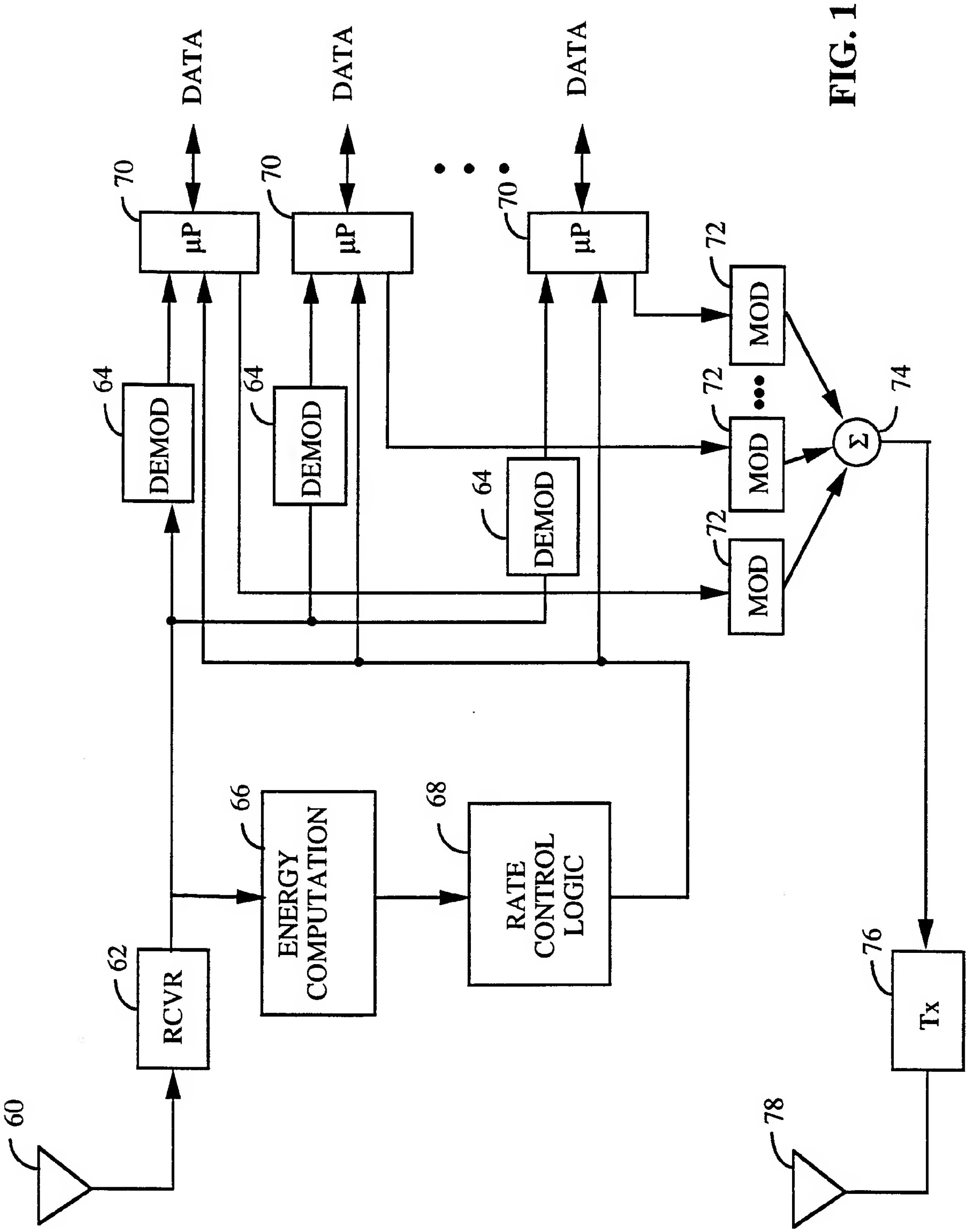
18
19
20

FIG. 10



DATA
DATA
DATA
70
70
70
μP
μP
μP
64
DEMOD
64
DEMOD
64
DEMOD
72
MOD
72
MOD
72
MOD
74
Σ
62
RCVR
66
ENERGY COMPUTATION
68
RATE CONTROL LOGIC
76
Tx
60
78

FIG. 11

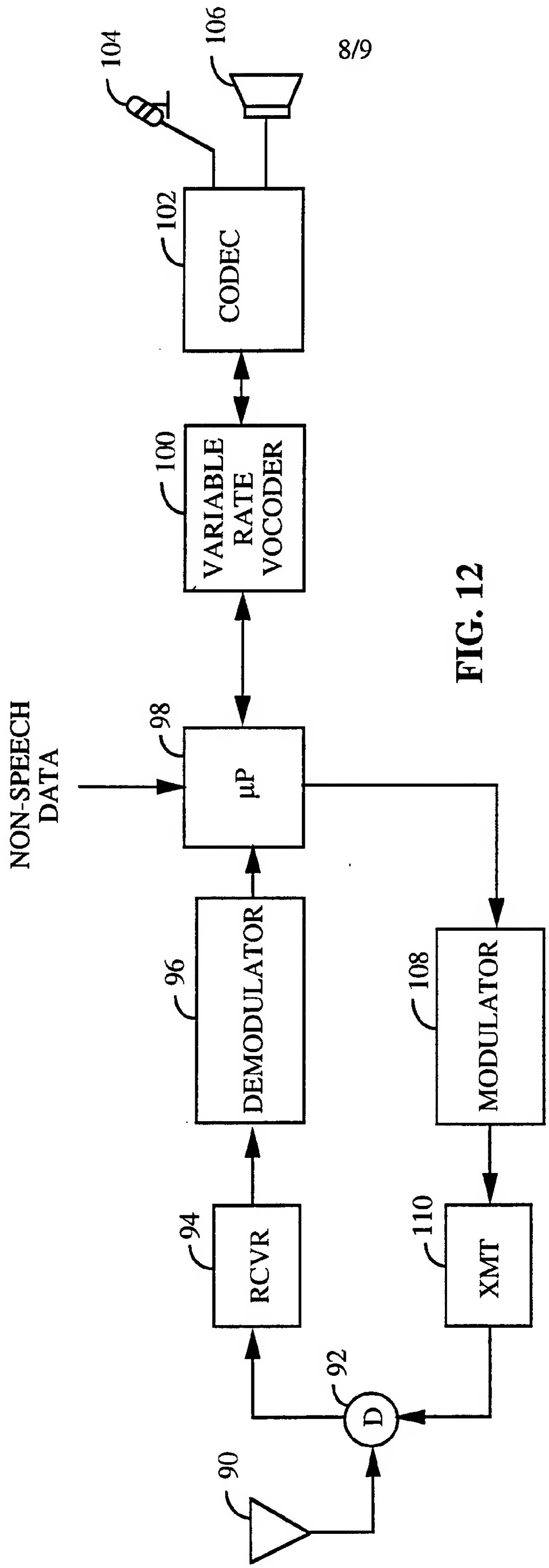
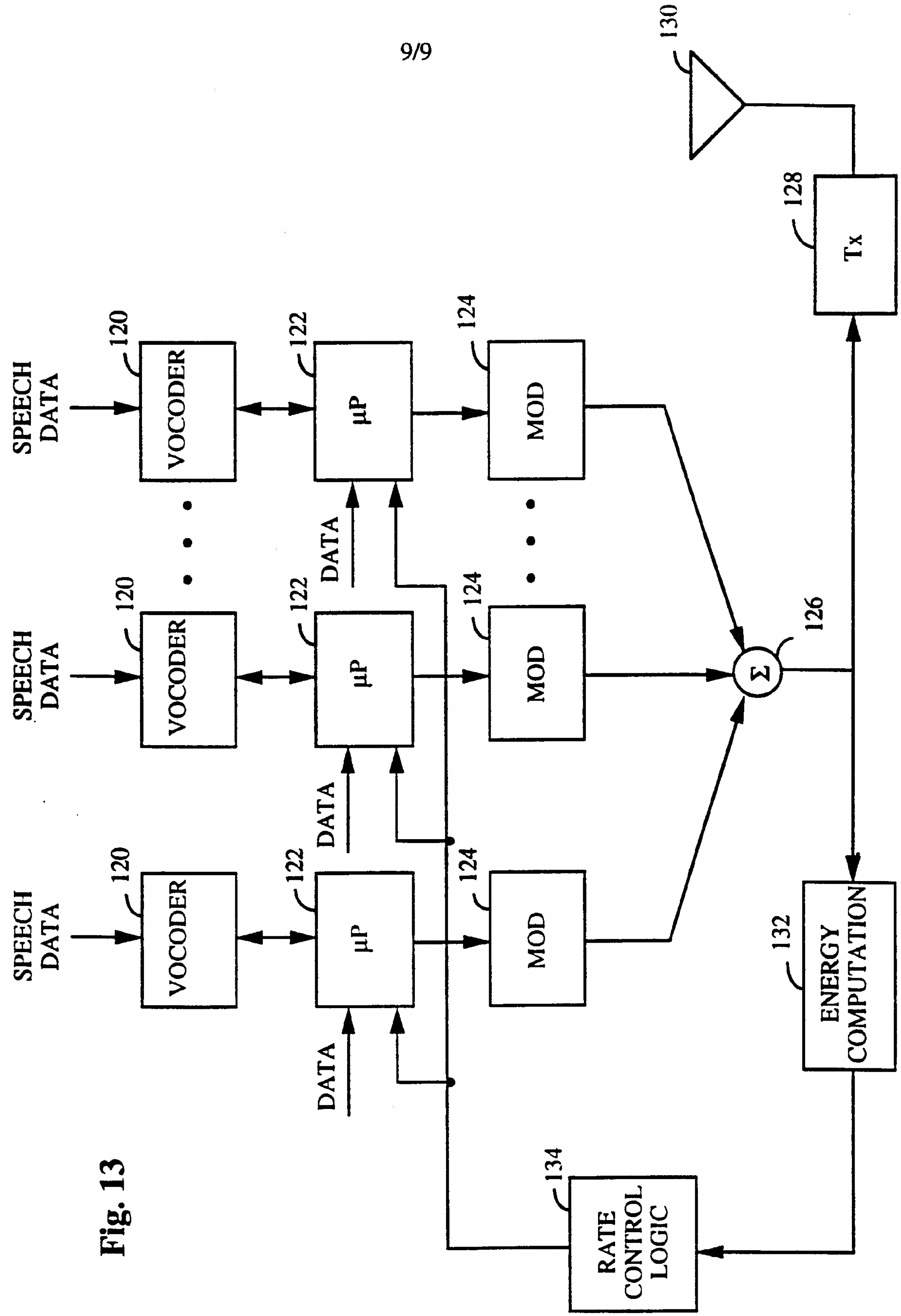

104
106
102
CODEC
100
VARIABLE
RATE
VOCODER
NON-SPEECH
DATA
98
μP
96
DEMODULATOR
108
MODULATOR
94
RCVR
110
XMT
92
D
90

FIG. 12



SPEECH DATA
VOCODER
120
µP
122
MOD
124
DATA
SPEECH DATA
VOCODER
120
µP
122
MOD
124
DATA
SPEECH DATA
VOCODER
120
µP
122
MOD
124
DATA
Σ
126
Tx
128
130
ENERGY COMPUTATION
132
RATE CONTROL LOGIC
134

Fig. 13

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 94/10087

A. CLASSIFICATION OF SUBJECT MATTER
IPC 6 H04B7/26 H04L1/12 H04Q7/38

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 6 H04Q H04B H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	EP,A,0 538 546 (MOTOROLA) 28 April 1993 see column 2, line 58 - line 21	1-20
X	EP,A,0 353 759 (NORAND CORPORATION) 7 February 1990 see column 2, line 29 - column 3, line 1	1,2,12, 13,15,18
X	EP,A,0 472 511 (ERICSSON) 26 February 1992 see column 3, line 37 - line 57	1,2,12, 13,15,18

[ ] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

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"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

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"&" document member of the same patent family

1

Date of the actual completion of the international search	Date of mailing of the international search report
3 November 1994	13. 12. 94

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+ 31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+ 31-70) 340-3016	Bischof, J-L

INTERNATIONAL SEARCH REPORT

Information on patent family members

Inter. ...onal Application No
PCT/US 94/10087

Patent document cited in search report	Publication date	Patent family member(s)		Publication date
EP-A-0538546	28-04-93	WO-A-	8706082	08-10-87
		AU-A-	5589086	20-10-87
		DE-D-	3689979	25-08-94
		EP-A-	0261112	30-03-88
		EP-A-	0412583	13-02-91
EP-A-0353759	07-02-90	US-A-	4910794	20-03-90
		AU-B-	632055	17-12-92
		AU-A-	3927889	08-02-90
		CA-A-	1316218	13-04-93
		GB-A,B	2223914	18-04-90
		US-A-	5070536	03-12-91
EP-A-0472511	26-02-92	AU-B-	642760	28-10-93
		AU-A-	8261991	27-02-92
		JP-A-	4234232	21-08-92
		NZ-A-	239283	27-09-94
		US-A-	5327576	05-07-94